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THE DESIGN OF ELECTRONIC MEASURING INSTRUMENTS.

By F. G. SPREADBURY, M.Inst.B.E., M.A.I.E.E.

Published by The Association of Engineering and Shipbuilding Draughtsmen,
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By F. G. SPREADBURY, M.Inst.B.E., M.A.I.E.E.

INTRODUCTION.

THE subject of Electronics is now approximately 50 years old and has touched and, in many instances, embraced nearly every field of science and technology, with beneficial results. Among the last of these fields is that of measurement. The delay here was probably due to the somewhat unreliable character of valves when applied to precision measurements, and resulted in the continued use of older types of instruments wherever these could be employed. However, it may be said that two early, highly important and outstanding instruments were born of sheer necessity and, despite the fact that now, many more electronic instruments have been added to their number, these two instruments are, in spite of their early origin, still those of greatest importance. These instruments are the cathode ray oscillograph and the valve voltmeter and partly owe their great importance to the fact that they now form essential parts of other instruments of greater complexity.

In recent years great improvements have been made in the uniformity and stability of valve characteristics, and this together with the discoveries of negative feedback and methods of supply voltage stabilization has resulted in the development of a very large number of different types of electronic instruments. These instruments have greatly extended the range of the older electrodynamic types and have made possible measurements hitherto impossible.

In a work of the present size it is, of course, impossible to describe all of the many instruments that have been devized. However, many of these instruments comprise a number of basic types and it is to these types that attention will be directed. In addition consideration will be given to valve characteristics relative to instrument design and the stabilization of power supplies. Also a number of actual instruments will be described.

Valve Characteristics and their Relation to Measurement.

With the assumption that electrons are emitted from a heated cathode with zero velocity, it may be shown* that the relation

* *Electronics*, F. G. Spreadbury (Pitman), p. 274.

between anode current and voltage in a vacuum diode is given by the law

$$I = KV^{3/2} \qquad (1)$$

where V is the sum of the applied anode voltage and contact potential difference between anode and cathode. A typical I/V characteristic is shown by Fig. 1, from which it is evident that the three-halves power law, or space-charge equation, as (1) is termed, does not hold over the entire characteristic. The curve may be roughly divided into five parts of which that lying between *c* and *d* may be considered to obey (1). Between *a* and *b* an exponential law is followed while between *b* and *c* the law is parabolic. *de* is linear but beyond *e* the curve bends over as saturation is approached.

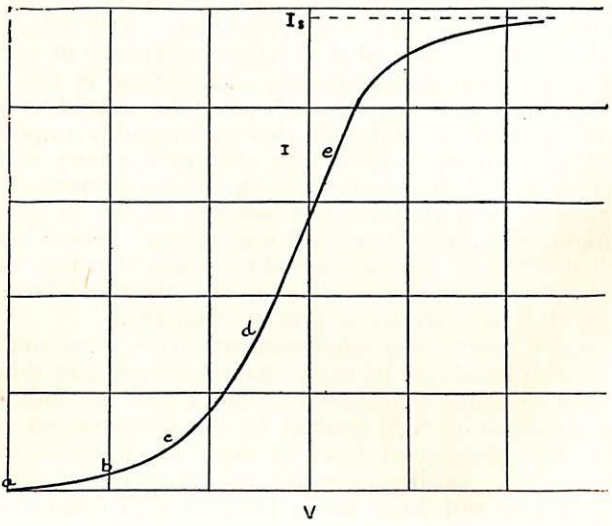


Fig. 1.

The cause of *ab* having an exponential form is that the electrons are not emitted with zero velocity from the cathode as assumed in deriving the space-charge equation. Actually the electrons are emitted with velocities obeying Maxwell's Velocity Distribution Law. This means that a continuous spectrum of velocities exists from zero to infinity. With the assistance of Maxwell's Law it may be shown* that the number of electrons emitted per second with energies lying between E and E + *d*E is

$$dn = \frac{n}{kT} \epsilon^{-E/kT} dE$$

* *Electronics*, p. 170.

where n is the number of electrons emitted per sq. cm. of cathode surface per second, T is the absolute temperature of the cathode and k is Boltmann's constant, equal to $1{\cdot}37 \times 10^{-16}$ erg/degree. If a potential V is applied to the anode then only those electrons with energies $E > -eV$ will cross the interelectrode space, where e is the numerical value of the electronic charge. Hence under these conditions the thermionic current is

$$i = \int_{-eV}^{\infty} edn = \frac{ne}{kT} \int_{-eV}^{\infty} \epsilon^{-E/kT} dE$$

or, $$i = ne\ \epsilon^{eV/kT}$$

But ne is the saturation current I_s. Hence

$$i = I_s\ \epsilon^{eV/kT} \qquad (2)$$

It must be remembered that (2) is correct only while V is a retarding potential, *i.e.*, while the potential of the anode is negative with respect to the cathode and space-charge effects are negligible. Obviously (2) is untrue for $V = 0$.

In many measuring instruments the no-signal electron current, given by (2), is very important because, usually, it is an unwanted quantity that means must be employed to suppress. Theoretically, for i to be zero $V = -\infty$. However, in practice, i can usually be brought sufficiently low by the application of a few volts and, furthermore, leakage paths have modifying effects.* It is apparent that variation of T will produce changes in i particularly as I_s is also a function of T. Thus

$$I_s = AT^2\epsilon^{-e\phi/kT}\ \dagger$$

where ϕ is the work function of the cathode material. The effect on i of varying the heater voltage of a diode is shown by Fig. 2, V, in this instance, being merely the anode-cathode contact potential.

As stated previously (2) is true only while the anode is negative to cathode. Taking logarithms of both sides of (2)

$$\log i = \log I_s + eV/kT$$

and, while (2) holds, $\log i$ is a linear function of V. Fig. 3 shows $\log i$ plotted against V for a 6H6 diode from which it is apparent that (2) holds up to about $-0{\cdot}6$ volt and 6·3 microamps.

As V is increased and space-charge effects become increasingly important (2) no longer holds and the characteristic passes through a transition stage which is parabolic. Normally, of course, the saturation stage is never approached because overloading would result in this region.

Triodes.

The anode current-grid voltage characteristic of a triode is similar to the amp-volt characteristic of a diode but the character-

* See p. 16.

† *Electronics*, p. 156.

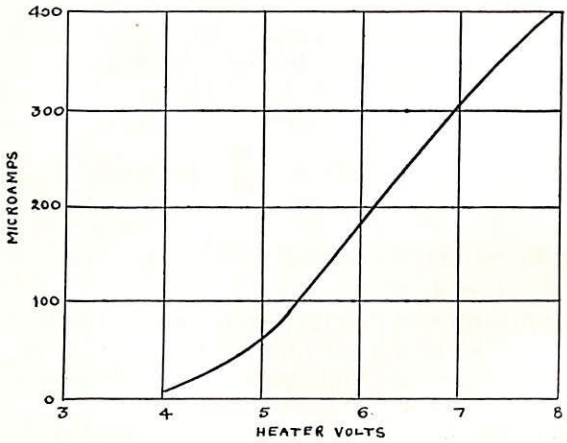

400
300
200
100
0
MICROAMPS
3
4
5
6
7
8
HEATER VOLTS

Fig. 2.

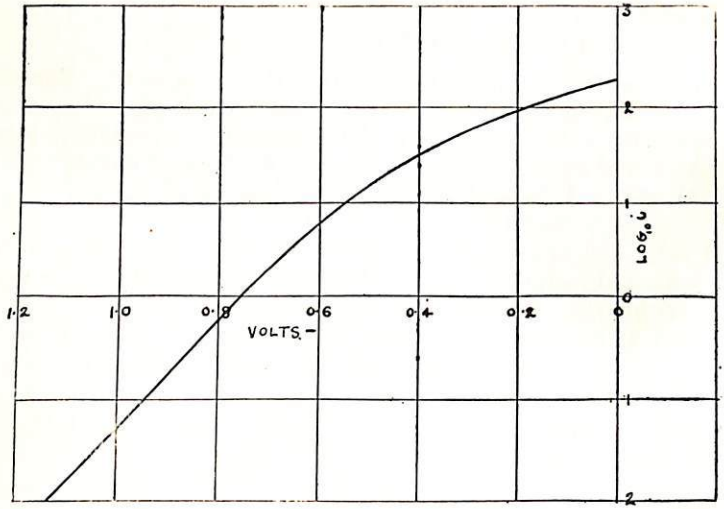

1·2
1·0
0·8
0·6
0·4
0·2
0
VOLTS
3
2
1
0
1
2
LOG

Fig. 3.

istics become shifted to the left as the anode voltage is increased. As with a diode, an exponential law is followed only while the grid voltage is negative, the magnitude of this voltage being, generally, considerably greater than that for a diode and increasing with an increase in anode volts. The parabolic section is also obtained with negative grid voltages, again shifting to the left with increasing anode voltage. At zero grid voltage the characteristic is usually linear while conditions for positive grid voltages are seldom of interest owing to the flow of grid current that occurs with the resultant lowering of input impedance.

The exponential characteristic is employed in slide-back voltmeters, while a parabolic characteristic is of fundamental importance in r.m.s. voltmeters, wattmeters, harmonic analysers, etc. A linear characteristic is desirable in d.c. voltmeters, such as those described on p. 28.

Input Impedance.

Among the outstanding advantages of many electronic measuring instruments is that of very high input impedance and consequent low loading of the source upon which measurements are made. The input impedance of a valve is dependent upon interelectrode capacitances, lead inductances, electron transit times, and electron and ion currents flowing between the electrodes and in associated leakage paths. The cathode-grid current in a triode is the sum of several currents that may be listed as follows : (*a*) The electron current previously discussed ; (*b*) a positive ion current due to residual gas within the valve ; (*c*) leakage currents across insulation materials ; (*d*) electron emission from the grid, and (*e*) currents due to photo-electric effects. Curves of all but the last of these currents and their resultant are shown by Fig. 4. It will be noted that all are functions of grid voltage and that the resultant is zero at P. Hence if operation can be confined to this point the input resistance will be a maximum. External currents on opposite

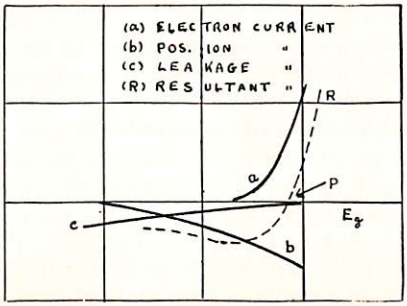


Fig. 4.

sides of P are, of course, oppositely directed. Provided the grid is negatively biassed the grid current is usually sufficiently small for most purposes. However, for measurement of extremely small currents, such as those occurring in ionization chambers, grid current must be reduced to a very low value. This is further discussed on p. 46.

Inter-electrode capacitance, lead inductance and electron-transit time are mainly of importance with high-frequency measurements. Evidently the direct effect of grid-cathode capacitance is to lower the input impedance of a valve. For ordinary triodes, and pentodes connected as triodes, $C_{g\text{-}k}$ is of the order of $5\mu\mu F$. However, in miniature sizes $C_{g\text{-}k}$ may be as low as $1-2\mu\mu F$. The foregoing capacitance is not the only one to be considered because the grid circuit is coupled to the anode circuit by the anode-grid capacitance $C_{a\text{-}g}$. When the anode load resistance is negligibly small (as it usually is with valve-voltmeters) it may be shown* that the input capacitance is

$$C_i = C_{g\text{-}k} + C_{a\text{-}g}$$

and, for a triode, $C_{a\text{-}g}$ may be as large as $C_{g\text{-}k}$. It is apparent that an input capacitance of $10\mu\mu F$. at radio frequencies may present a serious load to a source of negligible power and, in this respect, triodes are at a disadvantage. However, even with pentodes, where $C_{a\text{-}g}$ is negligible, C_i is frequently as high as $5\mu\mu F$.

The inter-electrode capacitance of a diode is usually less than that of triodes and pentodes and, because of this, diodes are employed where possible when measurements at high frequencies are made. Thus diodes are frequently employed in h.f. valve voltmeters. The values of $C_{a\text{-}k}$ for the EA 50 and EY 51 diodes are, respectively, 2·1 and $0{\cdot}8\mu\mu F$. Germanium diodes have even lower values, that of the GEX 55 being $0{\cdot}3-0{\cdot}7\mu\mu F$. It must be remembered, of course, that valve capacitances are invariably increased by associated stray capacitances in the circuit in which the valve is employed.

Inductance.

At very high frequencies the inductance of the leads connecting a valve to the source undergoing measurement may, in conjunction with the valve capacitance, cause resonance effects. This means that the voltage across the valve electrodes may be higher than that of the source and this, of course, will cause a voltmeter to read high. Furthermore, since it is a series resonant circuit that is formed, the input impedance falls as resonance is approached. As the connecting leads cannot be reduced indefinitely, it is the electrode and stray capacitances that finally limit the frequency at which an instrument such as a valve voltmeter may be employed. Hence

* *Electronics*, p. 320.

it is particularly desirable that these capacitances should be as small as possible. The employment of a diode within a probe, of course, represents an attempt to keep capacitances and lead inductance to a minimum. Probe details are given on p. 25. In order to form some idea of the effect of resonance, let C be the input capacitance of a valve voltmeter. Then at frequencies well below resonance the input reactance is $-1/pC$ ohms. Ignoring the effects of losses, the input impedance is $pL - 1/pC$, where L is the lead inductance and $p = 2\pi f$. Hence the ratio of the voltage on C to the input voltage is

$$\frac{1}{1 - p^2LC} = \frac{1}{1 - p^2/p_0^2}$$

$$= \frac{1}{1 - (f/f_0)^2}$$

where f and f_0 are, respectively, the operating and resonant frequencies. With $f < f_0$ this ratio is greater than 1 showing that the meter will read high. Hence the voltmeter readings should be multiplied by $1 - (f/f_0)^2$.

Electron Transit Time.

At very high frequencies, of the order of 10^7 c/s, the time for an electron to cross the interelectrode space may be comparable with that of a period of the alternating e.m.f. Under these circumstances the potential difference across, say, a diode, and, hence, the field between the electrodes, may reverse before the electrons reach the anode. When these effects are sufficiently pronounced electrons may fail to reach the anode, being reversed in the interelectrode space and returned to the cathode. Such phenomena are termed transit-time effects and with the diode voltmeter circuits of Fig, 6 result in the capacitors failing to charge to the peak value of the supply voltage. Hence transit-time effects cause such voltmeters to read low. Similar effects, of course, tend to occur with triodes and triode voltmeters. It is apparent that transit-time effects increase with frequency and decrease with a reduction in the interelectrode distance. They also increase as the magnitude of the applied voltage decreases because the electron accelerating force is proportional to this voltage.

Considering a voltmeter employing one of the circuits of Fig. 6, the instantaneous voltage across the diode is

$$v = V_m \sin pt - V_c$$

where V_c is the steady voltage to which C is charged. In the absence of loading and transit-time effects, $V_c = V_m$. Otherwise

$V_c < V_m$. Assuming a uniform field in the diode, the value of this is

$$X = v/d = (V_m \sin pt - V_c)/d$$

where d is the anode-cathode distance. The force accelerating the electron is

$$eX = md^2x/dt^2$$

$$\text{and } d^2x/dt^2 = e(V_m \sin pt - V_c)/dm \quad (3)$$

where e, m and x are respectively the electronic charge, mass and displacement. Evidently d^2x/dt^2 is positive while $V_m \sin pt > V_c$ and negative when the reverse obtains. The electron velocity is obtained by integrating (3). Assuming the initial electron velocities to be zero,

$$dx/dt = (e/pdm)\,[\,V_m(\cos pt_0 - \cos pt) + V_c(pt_0 - pt)\,]$$

where $pt_0 = \text{arc sin } V_c/V_m$. Integrating again

$$x = (e/p^2dm)\,[V_m(\sin pt_0 - \sin pt) - V_m \cos pt_0\,(pt_0 - pt) - (V_c/2)(pt_0 - pt)^2\,].$$

Under the influence of a sufficiently high frequency, $dx/dt = 0$ for $x \leq d$. Then

$$V_c/V_m = \sin pt_0 = (\cos pt_0 - \cos pt)\,/\,(pt - pt_0) \quad (4)$$

from which pt can be derived as a function of V_c/V_m. Also

$$f = \frac{\sqrt{V_m}}{2\pi d}\sqrt{\frac{e}{m}}\,[\,(\sin pt_0 - \sin pt) - \cos pt_0\,(pt_0 - pt) - \sin pt_0\,(pt_0 - pt)^2/2] \quad (5)$$

and substituting the values of pt and pt_0, derived from (4), in this expression gives the value of the frequency corresponding to a given value of V_c/V_m. For a given error it is apparent that $f \propto \sqrt{V_m}/d$.

The values of pt are plotted in Fig. 5 against V_c/V_m and the percentage error. The square root of the quantity in the square brackets in (5) is also plotted and will be noted as a straight line. Hence, on substitution, the percentage error is $7fd\sqrt{V_m}$ where f is the signal voltage frequency in M_c/s and V_m is in volts.

Stability.

It is evident that when valves are employed for measuring purposes the stability of their characteristics is of considerable importance. It is also important when one valve is replaced by another that calibration of the instrument is unaffected. Before employing a valve in a measuring instrument it is advisable to subject it to an ageing process in order to stabilize the emission and electrode contact potentials. This may take the form of running it for a period of about 100 hours. At the end of this period further variations in characteristics are likely to be far less

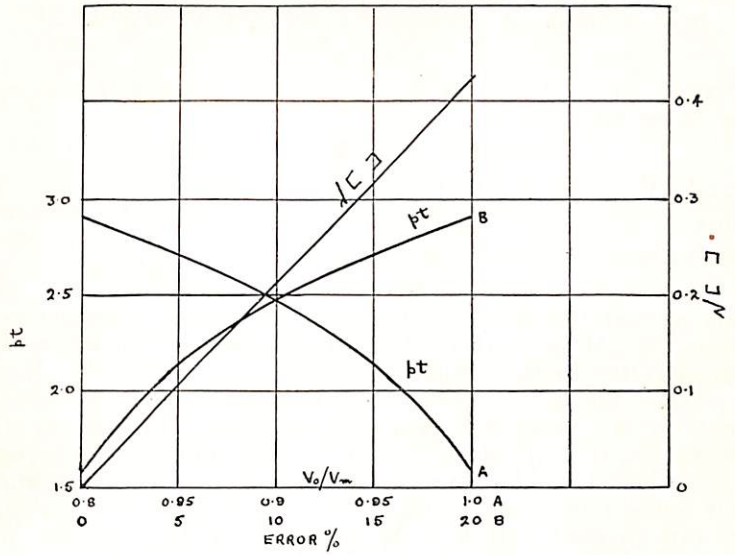


Fig. 5.

rapid than at the beginning. In instances where recalibration of an instrument in unavoidable on changing a valve, it is apparent that the valve should have as high an expectation of life as possible.* Valves with high expectations are available, *e.g.*, Brimar T valves.

Negative Feedback.

One of the most effective methods of rendering instrument accuracy independent of valve replacements and variation in valve characteristics is the employment of negative feedback. This is dealt with in specific instances later and here a general treatment only will be given.

Electronic measuring instruments frequently include some form of amplifier and it is to amplifiers that feedback is commonly applied. Let a voltage V_i be applied to the input terminals of an amplifier and an output voltage V_0 result. Following this, let a fraction α of the output voltage be fed back in such a manner that a voltage αV_0 is connected in series with V_i so as to reduce the input voltage. Then

$$V_{ir} = V_i + \alpha V_0 \qquad (6)$$

where V_{ir} is the resultant of V_i and αV_0. Vector notation must be assumed in this connection because V_i and V_0 are not necessarily

* *The Life of Oxide Cathodes in Modern Receiving Valves.* G. H. Metson, S. Wagener, M. F. Holmes and M. R. Child. J.I.E.E., Pt. III, Mar., 1952, p. 69.

in phase or antiphase. The gain of an amplifier is given by

$$M = V_0/V_{ir}$$

Hence $V_{ir} = V_0/M$

and substituting in (6)

$$V_0 (1/M - \alpha) = V_i$$

But V_0/V_i is the resultant gain, M_r.

Thus $M_r = M/(1 - \alpha M)$

and if αM is large compared with 1

$$M_r = -1/\alpha \qquad (7)$$

showing that the gain depends only on the system producing feedback. If $\alpha M \gg 1$, then $M_r \ll M$ and the gain is much smaller with negative feedback than when this is not employed. Hence to secure the advantages attached to feedback some loss of amplification must be sustained. However, turning to the advantages, it is apparent that if (7) holds and α is independent of frequency then the amplifier is free from frequency distortion. Now consider the case when a harmonic is produced by a valve in its output circuit. If A is the amplitude of this harmonic, this amplitude will appear in the output in the absence of feedback. However, when feedback is employed, the amplitude is, say, A_1. The amount fed back is αA_1 and this is amplified M times. Hence the component in the output due to A_1 is $\alpha M A_1$. In addition there exists the component A and it follows that the vector sum of these two components must be equal to A_1. Hence

$$A_1 = A + \alpha M A_1$$

$$\text{or } A_1 = A/(1 - \alpha M)$$

This result shows that harmonic distortion may be reduced by the employment of negative feedback. Thus, if $|1 - \alpha M|$ is 10, harmonic distortion is only one tenth of the value it would have in the absence of feedback. Because the magnitude of the harmonic depends on the output voltage, the latter must be the same in both cases, *i.e.*, with and without feedback. This means that V_i must be $|1 - \alpha M|$ times as great in the presence of feedback as in its absence. Of course, when feedback is negative, α is a negative quantity.

Output Impedance.

By means of what is termed voltage feedback the slope resistance of a valve may be reduced and, among other changes, the output impedance decreased thereby. This is because a fall in output voltage reduces the amount of feedback. Let Z_0 be the output impedance of an amplifier in the absence of feedback and I_0 the output current. Then the fall in output voltage is $I_0 Z_0$ and this fall may be regarded as introduced by the valve. With feedback,

the effective output impedance is Z_0', and the voltage fed back due to I_0Z_0' is $\alpha I_0Z_0'$. This is amplified to appear in the output as a component $\alpha MI_0Z_0'$. Hence there are two component voltage changes in the output and the vector sum of these must be equal to I_0Z_0'.

Hence $$I_0Z_0 + \alpha MI_0Z_0' = I_0Z_0'$$
and $$Z_0' = Z_0/(1 - \alpha M)$$

An alternative method of deriving this result is the following. From (6)

$$V_{ir} = V_i + \alpha V_0$$

Multiplying by M

$$MV_{ir} = MV_i + \alpha MV_0$$
or $$V_0 = MV_i + \alpha MV_0$$

Differentiating with respect to I_0, the output current,

$$\frac{dV_0}{dI_0} = \frac{d\,(MV_i)}{dI_0} + \alpha M \frac{dV_0}{dI_0}$$

But $dV_0/dI_0 = Z'_0$ and $d\,(MV_i)/dI_0 = Z_0$

Hence

$$Z_0' = Z_0 + \alpha MZ_0'$$
and $$Z_0' = Z_0/(1 - \alpha M)$$

Returning to a consideration of (7), this result was obtained on the assumption that $\alpha M \gg 1$. When this is not so, let a change $\triangle M$ result in a change $\triangle M_r$. Then

$$M_r + \triangle M_r = (M + \triangle M) / [1 - \alpha(M + \triangle M)]$$

Dividing throughout by M_r

$$1 + \triangle M_r/M_r = (M + \triangle M) / [1 - \alpha(M + \triangle M)]\, M_r$$

and, after simplification, this becomes

$$\triangle M_r/M_r = \triangle M/M\, [1 - \alpha(M + \triangle M)]$$

or, when $\triangle M$ is indefinitely small,

$$\frac{dM_r/M_r}{dM/M} = \frac{1}{1 - \alpha M}$$

Thus if $|1 - \alpha M| = 10$, a 10 per cent. change in M results in a 1 per cent. change in M_r.

Valve Voltmeters.

Possibly the most important electronic measuring instrument after the oscillograph is the valve voltmeter. This is because this instrument not only performs direct measurements that can scarcely be carried out in any other way, but because it also forms an essential part of many other instruments, such as timers, light-meters, harmonic analysers, fluxmeters,* etc. The unique properties of the valve voltmeter are its extremely high input impedance

* *Electronic Fluxmeter*, P. L. Taylor, *Wireless World*, April, 1951, p. 161.

and independence of frequency over a very wide range. With regard to the latter property, single instruments may be constructed that are consistently accurate over a range extending from about 20 c/s to 100 Mc/s. Valve voltmeters may be employed for either d.c. or a.c. measurements but it is in the a.c. field that these instruments find their most numerous applications. An instrument employing a single valve only seldom measures voltages less than 0·1 volt but this limitation is easily removed by means of pre-amplification. A large number of voltmeters have been devised but these may be divided into a relatively few types depending on their applications and method of operation. Thus the three outstanding classes of valve voltmeters are those respectively measuring average, r.m.s., or peak values. Most instruments employ hard valves and make use of rectification. However, a few instruments employ thyratrons and, sometimes, where high input impedance only is required, valves are employed mainly to endow an ordinary electrodynamic instrument with this property. Of course, with valves employed in this manner frequency limitations are mainly those of the instrument employed.*

Diode Voltmeters.

The simplest of all forms of valve voltmeters is the diode type, whether employed for measuring peak or average values. Reference to Fig. 6, (*a*) and (*b*), shows two methods of employing a diode for peak voltage measurements, while at (*c*) and (*d*) are shown two modifications of these. That of (*c*) is termed a shunt-type circuit and that of (*d*) a series type. With both (*a*) and (*b*) the capacitor tends to become charged to the peak value of the supply voltage, successively smaller increments of charge being acquired during each conducting period.† Theoretically, an infinite time is required for the capacitor voltage to become equal to the peak value of the supply but the actual time for the voltage to become steady depends on the impedance of the diode, the waveform and frequency of the supply and the leakage resistance of the system. With impulse voltages and large capacitors as long as half-a-minute may be required before the average value of the charging current declines to the value at which it just balances the leakage. It has already been shown that for small currents, the diode amp/volt characteristic is exponential. Hence even in the absence of a signal, provided the input terminals are connected, the capacitors in Fig. 6 should charge, theoretically, to an infinite voltage taking an infinite time in the process. However, owing to leakage, this does not occur and the capacitor

* *A Very High Impedance R.M.S. Voltmeter for Iron Testing*, D. C. Gall and F. C. Widdis, J.Sci.Insts., December, 1946, p. 287.

† *Capacitor Charging from Rectified Sources*, F. G. Spreadbury, *The British Engineer*, March, 1949.

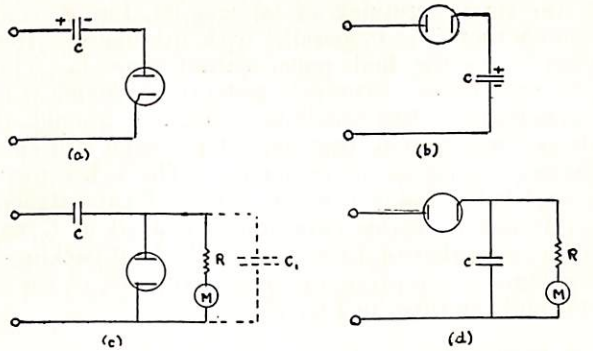


Fig. 6.

merely charges to a small finite voltage of the order of one volt. Let R_1 and R_2 be the respective values of the diode and capacitor leakage resistances. Then the total leakage current is

$$I = \frac{V}{R_1} + \frac{V}{R_2}$$

$$= \frac{V}{R} = I_s \epsilon^{eV/kT}$$

where $R = R_1 R_2 / (R_1 + R_2)$

Hence the emission current declines to the value at which it just equals the leakage current created by V. Since V is about 1 volt, the time to reach this voltage for capacitors of usually-employed values is so small that, for most purposes, the charging time may be regarded as negligible. Table 1 gives some values of V for different valves, taken by the author.

Table 1.

Valve	*V. Volts*
HVR2	0·888
EA50	1·132
EY51 (1)	1·38
EY51 (2)	1·27
EF37	0·935
354V	1·11

The last two valves were converted to diodes by connecting the grids to the anodes.

With the input terminals of (*a*) and (*b*), Fig. 6, connected it will be noted that C is in parallel with the diode. Hence there is a voltage across the diode equal to that across C. This voltage is often referred to as a "contact" potential although it is not due to the junction of two dissimilar metals. Should the input terminals be unconnected, this contact potential still exists across the diode and also across the terminals. This is because the latter may be regarded as constituting a very small capacitance in series with C. Hence it is this capacitance instead of C which then acquires the potentials of Table 1. Methods of backing-out diode contact potentials are given in the descriptions of the voltmeter of Fig. 17, and the timer of Fig. 47.

Effects of Heater Temperature.

The results given by Table 1 were obtained at normal heater voltages. Since the emission velocities are a function of temperature it is to be expected that the value of V will be dependent on the value of the heater voltage. That this is so is shown by Fig. 7 which concerns an EY51 diode.

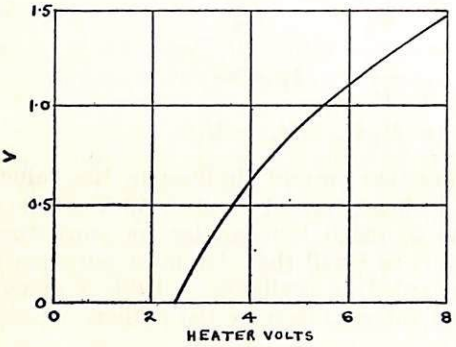


Fig. 7.

It was stated previously that with circuits of the type shown by Fig. 6, employed for voltage measurement, the capacitor tends to become charged to the peak value of the supply. Owing to the effect of the initial electron velocities, it is evident that a voltage in excess of the peak value is obtained, the amount of excess being equal to V volts. Hence, for low voltage measurements, considerable inaccuracy will result if the voltage across the capacitor is taken as equal to the peak value of the voltage being measured. Hence, either allowance must be made for the effect of initial electron velocities or the circuits must be restricted to measurements where 1 volt is negligible compared with the value of the voltage undergoing measurement. For voltages of about 100 volts and upwards the voltage on C may be measured

by means of an electrostatic voltmeter. A disadvantage of this, however, is that the voltmeter does not readily follow changes in the input voltage. It may be made to do so by shunting the voltmeter with a high resistance, although this may cause the meter to read low. Other simple methods of measuring the capacitor or diode voltage are indicated by the shunt circuits on C and the diode. These consist of a low-reading microammeter in series with a high resistance. This series combination constitutes, of course, a high resistance voltmeter. It causes the voltage on the capacitor to be lower than the peak value of that of the supply owing to the charge leaking away during the non-conducting half-cycles. Further methods of measuring the voltage on C will be discussed later in conjunction with other forms of voltmeters.

Efficiency of Rectification. Values of C and R.

This quantity is usually defined as the ratio of the d.c. voltage on C to the peak value of the a.c. supply voltage. It is a function of C and R declining with a decrease in the values of these quantities. C should be large in comparison with the diode capacitance in order that practically all the a.c. voltage is developed across the diode. Apart from this the reactance of C should be low compared with the resistance of R at the lowest frequency that will be experienced. As a provisional rule it may be stated that

$$R > 100/2\pi fC$$

$$\text{or } RCf > 100/2\pi$$

where f is the frequency of the voltage to be measured. The variation in rectification ratio with variation in R is shown by Fig. 8 for an EY51 at 50 c/s, the value of C being $0{\cdot}1\mu F$, the circuit being of the series type.

Effect of Low Voltages.

Assuming that compensation is made for the effect of contact potential, at relatively large input voltages the voltage across C is proportional to such voltages. However, below a few volts the relationship is non-linear owing to the relatively poor conductivity of the diode at low voltages. Thus, when conducting, the diode has difficulty in replacing the charge lost by C during non-conduction with the result that the voltage on C declines from the theoretically possible maximum to a greater extent than when the conductivity of the diode is relatively high. This effect is apparent from Fig. 8 and, as shown, must clearly increase as R is decreased.

Effect of R on Contact Potential.

Referring to (*c*), Fig. 6, it is apparent that the value of R will affect the value of the contact potential. Thus with R equal to

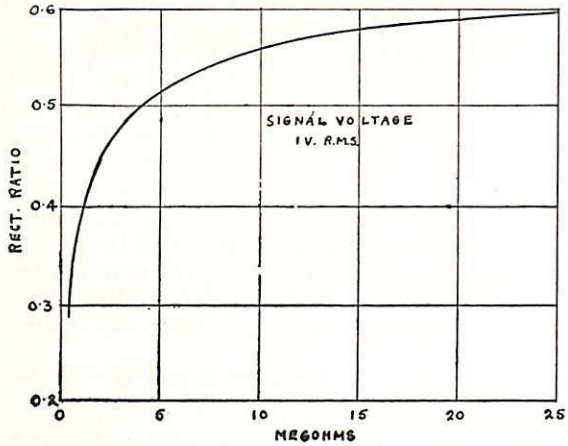


Fig. 8.

zero the no-signal current is a maximum and the contact potential zero. As R is increased the current decreases and the potential rises, the current becoming a minimum and the potential about 1·0 volt for R equal to infinity. Provided the input terminals are connected, similar effects occur with circuit (*d*).

Input Impedance.

At frequencies at which the capacitances of C and the diode can be neglected the circuits of Fig. 6 may be regarded as behaving as pure resistances. Referring to (*c*), let $V_m \sin pt$ be the value of the input voltage. Then if C charges to V_m the voltage on R is $V_m + V_m \sin pt$ or $V_m (1 + \sin pt)$. Thus the voltage on R is unidirectional but fluctuates between $2V_m$ and zero. The mean value of $V_m (1 + \sin pt)$ is, of course, V_m. The r.m.s. value of the voltage on R is

$$V_m \sqrt{\frac{p}{2\pi}} \int_0^{2\pi/p} (1 + 2 \sin pt + \sin^2 pt)\, dt$$

$$= \sqrt{\frac{3}{2}}\, V_m \qquad (8)$$

The energy loss in R is $3V_m^2/2R$, and if R_i is the effective input resistance

$$\left(\frac{V_m}{\sqrt{2}}\right)^2 / R_i = \frac{3V_m^2}{2R}$$

or $R_i = R/3$ (9)

For circuit (*d*)

$$\left(\frac{V_m}{\sqrt{2}}\right)^2 / R_i = \frac{V_m^2}{R}$$

$$\text{or } R_i = R/2 \qquad (10)$$

From these results it is apparent that for a high input impedance R should be as high as possible. However, the higher R becomes the lower is the sensitivity of the metering circuit. In practice the problem of low sensitivity with high values of R is overcome by amplifying the diode output. Suitable amplifying circuits are described on p. 28.

At very low and very high frequencies the capacitance of C and that of the diode affect the input impedance. The effect of C has already been discussed. In (*c*) the reactance of the diode capacitance acts as a shunt on R and thus lowers the input impedance. It is apparent that the effect of the diode capacitance can be serious because even if this is as low as $1\mu\mu F$ the reactance at 1 Mc/s is only $10^{12}/2\pi 10^6 = 0{\cdot}32 \times 10^6$ ohm. In practice associated stray capacitances render the input reactance even lower.

Since the capacitance across the diode is sometimes large, *e.g.*, in the wavemeter circuit of Fig. 32, further effects of this capacitance, apart from its lowering the input impedance, will be considered. Regarding Fig. 6 (*c*), it will be assumed that R is sufficiently large to be without effect, that the diode has negligible slope resistance, and that C_1 is the diode shunt capacitance. Then experiment supports the following statements :—

(1) C is charged to the peak value of the signal voltage once in each cycle.
(2) After the initial charge no current flows through the diode.
(3) If E and E_1 are the mean voltages on C and C_1, $(E + E_1) = 0$,
(4) If V and V_1 are the r.m.s. values of the alternating components of the voltages on C and C_1, $(V + V_1) = V_s$, where V_s is the signal voltage.

The current flowing in the capacitances is

$$I = pCV = pC_1V_1$$

$$\text{and } V_1 = VC / C_1$$

$$\text{or } V = V_sC_1 / (C + C_1)$$

$$\text{Also } V_1 = V_sC / (C + C_1)$$

$$\text{Again } E = \sqrt{2}V_s - \sqrt{2}V_sC_1 / (C + C_1)$$

$$= \frac{\sqrt{2}V_sC}{C + C_1} = -E_1$$

From these results it will be seen that effects of C_1 are to lower E and E_1, to cause voltage fluctuations on C and to reduce the amplitude of the fluctuations across the diode. Of course, nor-

mally the value of C_1 is so small compared with that of C that these effects are negligible. However, the current, given by

$$I = pCC_1V_s / (C + C_1) \fallingdotseq pC_1V_s$$

may be inconveniently large at high frequencies, as indicated when deriving the capacitive reactance above.

Comparison of Shunt and Series Circuits.

Considering circuits (*a*) and (*b*) it will be noted that the fundamental difference is that with (*a*) the mean voltage across the diode is measured while with (*b*) the voltage on C is measured. In practice, for valve voltmeter work (*a*) is commonly employed because C blocks any d.c. component in the signal voltage. Evidently if a d.c. voltage is applied to (*b*), with the upper terminal positive, C will charge to this voltage. If a series capacitor is employed to block the voltage it is ineffective because the latter then divides between C and the series capacitor. This also occurs when an a.c. signal is applied. Furthermore removal of the signal does not immediately remove the voltage from C. From (9) and (10), it is evident that, for the same value of R, the shunt circuit imposes heavier loading than does the series circuit.

When the heater of a diode in the series circuit is heated by means of a mains-fed transformer a problem may arise owing to the interwinding capacitance of this. Referring to Fig. 6, (*d*), let it be supposed that the lower terminal is at earth potential and that the heater and cathode are internally connected. Then with one side of the primary of the heater transformer connected to earth the live side of this is coupled to the cathode *via* the interwinding capacitance, C_w. Hence a relatively large a.c. signal voltage, usually 200-250, is applied between earth and cathode *via* C_w. Now if the normal input terminals are closed by means of the output impedance of some circuit of which the voltage is normally desired to be measured, then, assuming for the moment that this voltage is zero, the diode circuit virtually becomes that of Fig. 6, (*c*), (with the diode inverted) if it is assumed that the output impedance constitutes a short-circuit of the normal input terminals. The reader may convince himself of this by redrawing (*d*) with the terminals short-circuited and a signal impressed between earth and cathode *via* C_w at the cathode end. Turning to the now-equivalent circuit (*c*), the terminals here constitute the position of application of the previously-mentioned 200-250 volt signal, $C = C_w$ and C_1 corresponds to C in (*d*). Hence, from p. 21, across C_1 [or C in (*d*)], there is an alternating component

$$VC_w/(C_w + C_1)$$

and a d.c. component

$$E = \sqrt{2}VC_w/(C_w + C_1)$$

where V is the value of the mains supply voltage. Now if V = 240,

C_w = 0·0001μF and C_1 = 0·1μF, the values of these components are, respectively, 0·24 and 0·34 volt and these magnitudes would be quite serious in a low-reading voltmeter. Thus if the series circuit of (*d*), Fig. 6, were coupled to any of those of Fig. 12, closing of the diode input terminals by any means would immediately apply the above voltages to the voltmeter circuit.

The obvious solution of the foregoing problem is to have a low C_w/C_1 ratio. Usually C_1 is restricted to about 0·1μF, or less, and hence C_w should be as low as possible. This can be effected by means of an earthed metallic screen, such as metal foil, between the transformer windings. The screening should be as complete as possible, the foil completely enclosing one winding, care being taken, of course, to avoid any type of screen that virtually constitutes a short-circuited secondary. The effect of the screen is to reduce C_w to zero.

When the heater and cathode are not connected the foregoing effects are smaller and those that do occur depend on the impedance between the heater and cathode. However, for the measurement of high voltages it is impracticable to have the heater and cathode separate because the potential difference that the insulation between these will withstand is limited. In some small valves it is no more than 50 volts.

To demonstrate the effect of C_w the author employed the circuit of Fig. 6, (*d*), coupled to an amplifier of the type shown by Fig. 12, (*b*). A screened heater transformer was employed with an EY51 diode, the values of C and R_g being 0·1μF and 10M. Capacitance was then introduced between the transformer windings with the following results :—

$C_w\mu\mu F$	E_{cal}	E_{ob}
0	0	0
30	0·102	0·12
42	0·143	0·2
70	0·24	0·31
90	0·305	0·35
120	0·41	0·43
161	0·55	0·52
205	0·7	0·62
252	0·855	0·76

The transformer supply voltage was 240 a.c. and the calculated values were obtained from the second of the two equations above. By substituting capacitance to produce the same voltage as that introduced by the transformer without screening, the interwinding capacitance was found to be 50$\mu\mu$F.

It may be mentioned that the foregoing phenomena do not occur if the diode is reversed, *i.e.*, if the voltage to be measured is

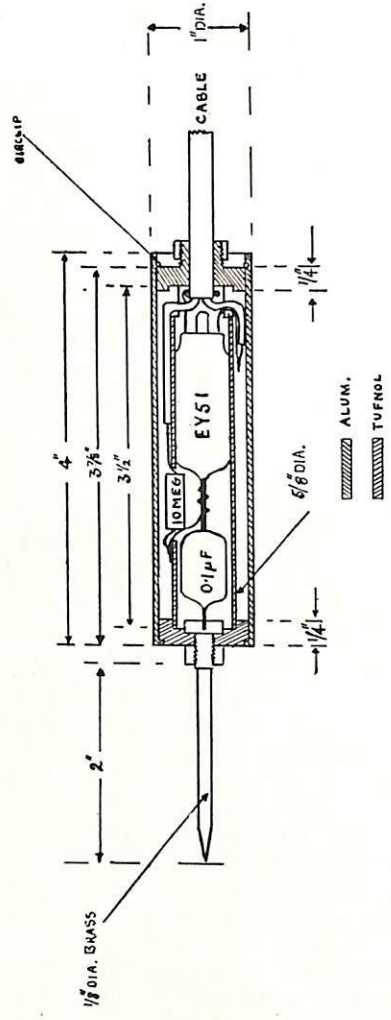

1" DIA.
CABLE
BIRCLIP
1/4"
EY51
10MEG
0·1μF
5/8" DIA.
4"
3 7/8"
3 1/2"
2"
1/4"
1/8" DIA. BRASS
ALUM.
TUFNOL

Fig. 9.

applied to the cathode instead of the anode. However, this brings the heater transformer into contact with the input terminals, either through interwinding or winding to earth capacitance.

Probes.

In making measurements at high frequencies it is evident that capacitance (and inductance) in the input circuit should be as small as possible. This requirement may be largely achieved by means of a probe. A probe, of the author's design, is shown by Fig. 9 and from this it will be seen that C and a miniature diode are housed in a tubular metal case from which protrudes a short pointed metal rod. The rod is directly connected to C and thus rectification may be effected very close to the point at which measurement is made. The leads from the probe to the instrument then carry only the rectified voltage and the 50 c/s supply for the diode heater. By means of a probe the input capacitance can be brought to a very low figure, that for the probe illustrated being $5\mu\mu F$.

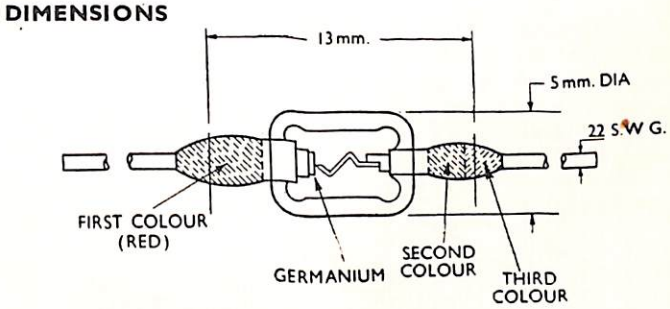
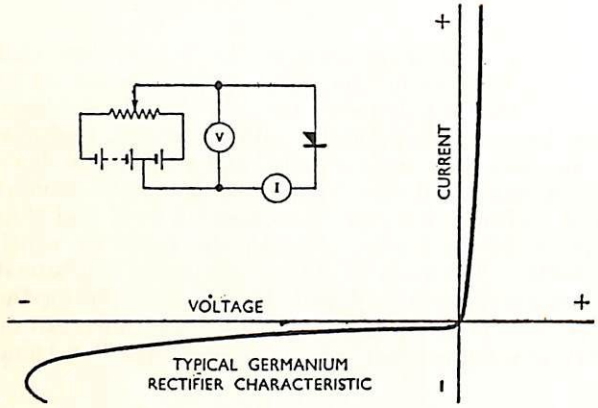


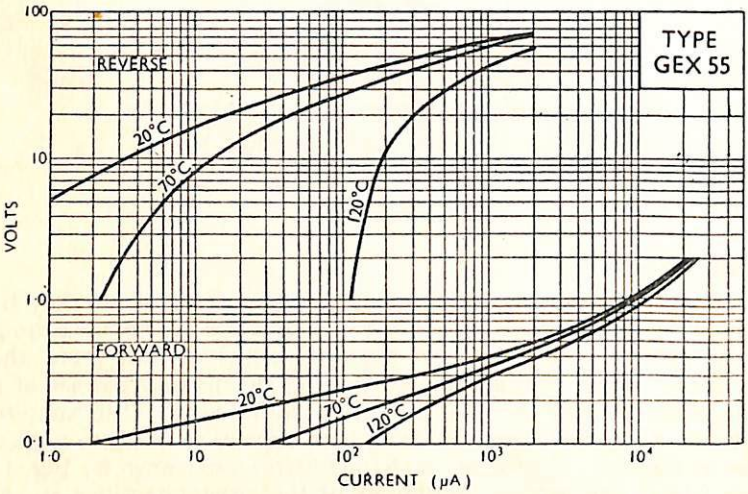
Fig. 10.

Crystal Diodes.

As an alternative to a thermionic diode a crystal diode may be employed.* Provided only small voltages are to be measured, a far smaller probe may be employed as is evident from the dimensions given by Fig. 10. Crystal diodes usually consist of a germanium crystal in contact with a spring-loaded tungsten or platinum wire, this assembly being hermetically sealed into a glass tube as shown. Rectification characteristics are shown by Fig. 11 from which two serious limitations of the crystal rectifier are at once apparent. These are the relatively low reverse resistance and voltage. It is evident that the reverse resistance progressively decreases as the voltage increases, becoming negative at a point known as the "turnover" voltage. Evidently the input voltage

* *Crystal Valves, Electrical Review*, 28th March, 1949, p. 507.





By courtesy of G.E.C., Ltd.

Fig. 11.

should not be such that this point is approached. Since the maximum reverse voltage across the rectifier is $2\sqrt{2}V$, we must have

$$2\sqrt{2}\,V < V_t$$

or

$$V < V_t / 2\sqrt{2}$$

where V_t is the turnover voltage. Thus if $V_t = 75$, $V < 26$ volts. Higher input voltages may be employed provided an attenuator is used. However, for h.f. work, this must precede the probe and thus increases the probe size.

In addition to its small dimensions, other advantages of the germanium diode are the absence of a heater and low capacitance. The value of the capacitance is about $0{\cdot}7\mu\mu F$ and this, of course, renders this diode advantageous for very high frequency measurements. Thus crystal diode voltmeters may be employed at at least 1000 Mc/s. The absence of a heater results in an absence of contact potential and hence no backing e.m.f. is needed.

Average-Reading Voltmeters.

The foregoing types of diode voltmeters read peak values. By removing the capacitor in (*d*), Fig. 6, a meter for reading average values is produced. With R equal to zero, the amp/volt characteristic obeys the three-halves power law and hence is non-linear. By employing high values of R the characteristic is linearized, the input impedance increased, and the calibration is rendered independent of variations in the diode characteristics. These advantages are, of course, secured at the expense of reduced sensitivity. If the diode impedance can be neglected, then with a sinusoidal input voltage, the instantaneous value of the current is

$$i = V_m \sin pt/R$$

and the average value

$$I_{av} = \frac{pV_m}{2\pi R} \int_0^{\pi/p} \sin pt\, dt$$

$$= V_m/\pi R \text{ amp.}$$

Hence $I_{av} \propto V_m/\pi$ and a moving coil meter may be calibrated directly in volts. However, whether the input is sinusoidal or not the meter will always read the average value of the current which will, in turn be proportional to the mean value of the input voltage.

Owing to their limitations and the fact that average values are seldom required, diode voltmeters of this type find limited uses. However, they are often used when preceded by an amplifier, and further details of this are given on p. 44.

Rectifier-Amplifier Types.

This type of instrument is virtually a peak voltmeter, although it is invariably calibrated as though reading r.m.s. values. As its name implies, it comprises a rectifier followed by an amplifier, the amplifier actually constituting a *d.c.* voltmeter and often being employed as such. The rectifier consists of either one of the two diode circuits shown by Fig. 6 or, alternatively, a crystal rectifier of the germanium type. After rectification, the output is passed to a d.c. amplifier which frequently takes the form of a bridge circuit. Typical basic amplifier circuits and their equivalents are shown by Fig. 12, while Figs. 18 and 47 show complete voltmeter circuits respectively embodying amplifiers of the types shown by (*d*) and (*b*). The amplifiers shown by (*c*) and (*d*) are known as difference amplifiers and it may be noted that (*b*) and (*d*) employ cathode-follower circuits. Since the circuits of Fig. 12 are operated by d.c. potentials, consideration of this fact must be taken in circuit analyses. Stability rather than sensitivity is of first importance in voltmeter circuits and experience shows that circuits (*b*) and (*d*) are more stable than those of (*a*) and (*c*). Because of this analyses will be made of (*b*) and (*d*) only. If desired, analyses of the other circuits may readily be made in a similar manner to that given below.

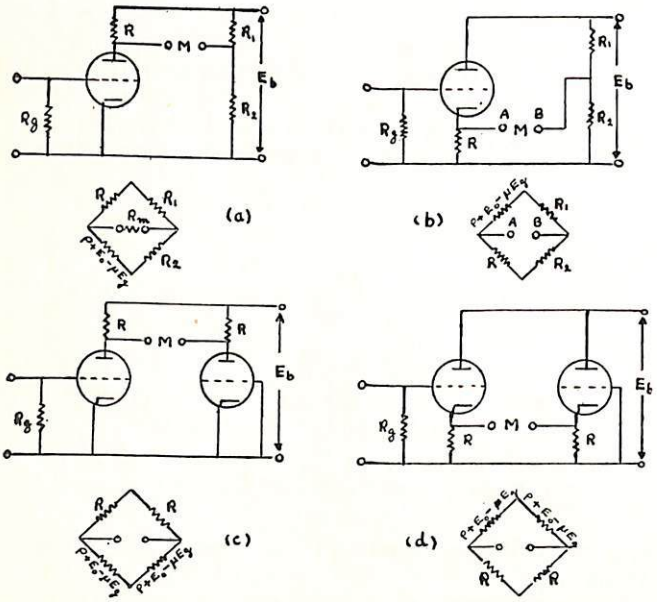


Fig. 12.

Referring to Fig. 13, it will be noted that the valve anode characteristics may be regarded as equidistant, parallel, straight lines, provided the lower parts are ignored. Considering any line and its hypothetic continuation into the fourth quadrant, the equation to this line is

$$I_a = a + bE_a$$

where I_a and E_a are, respectively, the anode current and anode-cathode voltage and a and b are constants for any particular line; b is the slope of the line and is equal to $dI_a/dE_a = 1/\rho$ where ρ is the anode resistance. Hence b has the same value for all lines. a is, of course, the intercept of a line on the I_a axis and has a different value for each line. However, a may be expressed as a function of the grid potential, E_g, in the following manner.

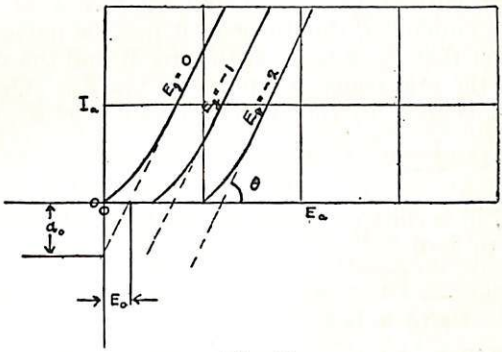


Fig. 13.

The value of E_a at any point where a particular line intercepts the OE_a axis is $(E_0 - \mu E_g)$ volts. Hence, since a is negative,

$$\frac{a}{E_0 - \mu E_g} = -\tan\theta = -\frac{1}{\rho}$$

$$\text{and} \quad a = \frac{\mu E_g - E_0}{\rho}$$

which leads to

$$I_a = \frac{\mu E_g - E_0 + E_a}{\rho} \qquad (11)$$

$$\text{or} \quad E_a = \rho I_a + E_0 - \mu E_g$$

Thus, in the equivalent circuit, the valve may be replaced by a resistance, ρ, in series with an e.m.f. $(E_0 - \mu E_g)$. Referring to Fig. 12 (b), the current in the left-hand arms is

$$I_a = \frac{\mu E_g - E_0 + E_b}{\rho + R}$$

or, since $E_g = -I_aR$,

$$I_a = \frac{E_b - E_0}{\rho + (\mu + 1)\,R}$$

If a voltage E_s is applied to the input terminals, then $E_g = E_s - I_aR$ and

$$I_a = \frac{E_b - E_0 + \mu E_s}{\rho + (\mu + 1)\,R}$$

Thus the voltage between A and B is

$$\frac{E_b - E_0 + \mu E_s}{\rho + (\mu + 1)\,R}\,R - \frac{R_2}{R_1 + R_2}\,E_b \qquad (12)$$

The meter current may be found by means of Thévenin's theorem. In employing this theorem it may be noted that, with the assumption that E_b is reduced to zero, R and the valve are in parallel with the grid connected to anode via R_g. Owing to this connection, I_a is given by (11) with E_g replaced by E_a, or

$$I_a = \frac{E_a\,(\mu + 1) - E_0}{\rho}$$

Hence, if, in accordance with the theorem, all e.m.f.'s. but E_a are reduced to zero.

$$I_a = \frac{E_a}{\rho/(\mu + 1)}$$

and the valve may be replaced by a resistance $\rho/(\mu + 1)$, this resistance being, incidentally, the output impedance of the cathode follower. Thus the meter current is

$$I_m = \frac{\dfrac{E_b - E_0 + \mu E_s}{\rho + (\mu + 1)\,R}\,R - \dfrac{R_2}{R_1 + R_2}\,E_b}{R_m + \dfrac{R_1 R_2}{R_1 + R_2} + \dfrac{R\,\rho/(\mu + 1)}{R + \rho/(\mu + 1)}} \qquad (13)$$

where R_m is the meter resistance.

For the circuit to be balanced the numerator must equal zero. Since the numerator contains E_b it is apparent that zero setting is dependent on the value of the H.T. voltage. Hence a stabilized supply is necessary for this type of circuit. With the circuit balanced in the absence of a signal

$$I_m = \frac{\mu R E_s}{R\rho + \left(R_m + \dfrac{R_1 R_2}{R_1 + R_2}\right)\left[\rho + (\mu + 1)\,R\right]} \qquad (14)$$

$$\text{or} \quad I_m = \frac{E_s}{\rho/\mu + \left(R_m + \frac{R_1 R_2}{R_1 + R_2}\right)\left[\frac{\rho}{\mu R} + 1 + 1/\mu\right]} \tag{15}$$

Thus

$$I_m \propto E_s$$

and, provided ρ/μ is relatively small, the circuit conditions are independent of the valve characteristics. Hence, for good stability, independency of ageing and valve replacements, R_m, R, R_1 and R_2 should be chosen so that the quantities ρ/μ and $\rho/\mu R$ are negligibly small. In making this choice the values of the various resistances must be made higher than if stability were a secondary consideration. An effect of this is reduced sensitivity, expressed as the ratio I_m/E_s. Thus reduced sensitivity is the price to be paid for increased stability, as already stated in dealing with negative feedback.

According to (15) I_m is proportional to E_s and is independent of the sign of E_s. However, these properties are realized only while the valve is operated on the straight portions of its characteristic. If R is too large, the relation between I_m and E_s is non-linear and positive values of E_s give larger values of I_m than corresponding negative ones. In fact, if R is very large negative values of E_s may cause cut-off. Hence attempts to obtain independence of the valve characteristics by making R large may lead to the above faults. In addition, if grid current is important, R must be chosen so that operation occurs to the left of the point P in Fig. 4.

In some instruments the sensitivity is controlled by the value R, or R_m, the value being increased with the range. In these instances it is usual to introduce a positive grid bias voltage in order partly to counteract the large negative bias produced by the drop on R. In this manner the net grid potential is held slightly negative to the cathode, causing the valve to work on the linear position of the characteristic. A result of this is that a much larger value of R may be employed than would otherwise be possible. This subject is further discussed on p. 33.

Comparing Fig. 12, (*d*), with Fig. 12, (*b*), it will be noted that the resistor R_1 is replaced by a valve and it will be assumed in the following analysis that both valves have identical characteristics. In the absence of R_m the currents in the left and right-hand arms of the bridge are, respectively,

$$I_{a1} = \frac{E_b - E_0 + \mu E_s}{\rho + (\mu + 1)\,R} \tag{16}$$

$$I_{a2} = \frac{E_b - E_0}{\rho + (\mu + 1)\,R} \tag{17}$$

and the voltage between A and B is $RI_{a_1} - RI_{a_2}$ or

$$\frac{\mu R E_s}{\rho + (\mu + 1)\, R} \tag{18}$$

From Thévenin's theorem, the meter current is

$$I_m = \frac{\dfrac{\mu R E_s}{\rho + (\mu + 1)\, R}}{R_m + \dfrac{2R\rho \,/\, (\mu + 1)}{R + \rho \,/\, (\mu + 1)}}$$

$$\text{or } I_m = \frac{E_s}{2\rho/\mu + R_m\,(\rho/\mu R + 1 + 1/\mu)} \tag{19}$$

This result shows that $I_m \propto E_s$ and that balance of the circuit is independent of the H.T. voltage, E_b. The latter characteristic is realized only if the valves have identical characteristics. Invariably the characteristics differ and in consequence the circuit balance is disturbed by variations in E_b. Furthermore, the circuit is not automatically balanced when $E_s = 0$. Although variations in E_b may cause less disturbance than with the circuit of Fig. 12, (*b*), it is usually necessary to stabilize the H.T. supply and to provide means for balancing. In Fig. 18, the latter is effected by means of R_{11}. If values are so chosen that ρ/μ and $\rho/\mu R$ are negligibly small

$$I_m \fallingdotseq \frac{E_s}{R_m} \tag{20}$$

and the circuit behaviour is independent of the valve characteristics. In these circumstances if R_m = 10K an input of 1 volt gives full scale deflection on a 100μA meter. The remarks made following the mathematical analysis of Fig. 12, (*b*), are also applicable to the present circuit. However, under the circumstances in which (20) was derived (15) becomes

$$I_m \fallingdotseq \frac{E_s}{R_m + \dfrac{R_1 R_2}{R_1 + R_2}}$$

Hence the sensitivity of two-valve circuit is the greater.

When the foregoing types of d.c. amplifiers are employed for a.c. measurements they are preceded by a rectifier, usually that of Fig. 6, (*a*). In order to avoid shunting the diode by the amplifier capacitance and to prevent a.c. reaching the amplifier, a filter, R_1C, is employed as shown by Fig. 14. This causes a substantially steady voltage to be presented to the amplifier terminals. However.

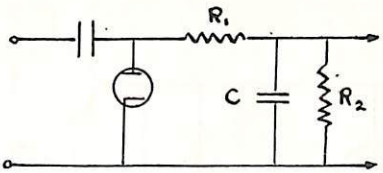


Fig. 14.

an effect of R_1 is to modify the diode input impedance and to reduce the sensitivity. Thus of a peak input voltage V_m, a fraction $R_2V_m/(R_1+R_2)$ is developed across C. If the filtering is such that the ripple on C is negligible then the voltage on R_1 consists of a d.c. component $R_1V_m/(R_1+R_2)$, say αV_m, and an alternating component $V_m \sin pt$. The r.m.s. value of this voltage is

$$V_m \sqrt{\frac{p}{2\pi}} \int_0^{2\pi/p} (\alpha^2 + 2\alpha \sin pt + \sin^2 pt)\, dt$$

$$= V_m (\alpha^2 + 1/2)^{\frac{1}{2}} \text{ volts.}$$

If the voltage on R_2 is written as βV_m, then the losses in the resistors R_1 and R_2 are

$$\frac{(\alpha^2 + 1/2)\, V_m^2}{R_1} + \frac{\beta^2 V_m^2}{R_2} \text{ watts}$$

and

$$\frac{V_m^2}{2R_i} = \frac{(\alpha^2 + 1/2)\, V_m^2}{R_1} + \frac{\beta^2 V_m^2}{R_2}$$

or

$$R_i = \frac{R_1 R_2}{2\,[\beta^2 R_1 + (\alpha^2 + 1/2) R_2]} \qquad (21)$$

where R_i is the diode input resistance.

Evidently for high input impedance R_1 should be high while for high sensitivity R_1 should be low. In practice a compromise must be effected between these conflicting requirements.

Rectifier-amplifier voltmeters principally consist of two classes, those similar to the example of Fig. 18, where the range is changed by means of the attenuator $R_2 - R_7$, and those represented by Fig. 15 where range-changing is effected by change of R or R_m or both. With both circuits the grid may be made positive by inverting the diode, always provided that the grid to cathode potential is maintained negative. With the circuit of Fig. 15, on the higher ranges relatively large voltages are set up on C and these in conjunction with the voltage drop on R would make the grid to cathode potential excessively negative. Hence the lower end of R is made positive to cathode in order to cause the working point of the triode to lie on the linear part of the characteristic. By means of ganged switches R and the bias are simultaneously changed in order that

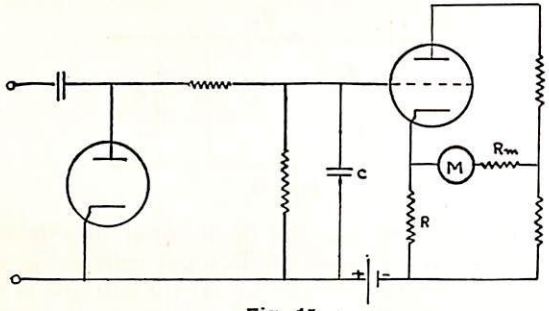


Fig. 15.

the working point is the same for all ranges. Of course, a high voltage signal produces a relatively small grid-cathode potential change because as the grid goes more negative the drop on R decreases.

Drift.

In discussing the foregoing voltmeter circuits, various limitations have been mentioned, and means of minimizing these. There is, however, a further limitation, and one that is common to most d.c. amplifiers, this limitation being termed *drift*. With reference to a voltmeter circuit, drift implies that subsequent to balancing the circuit, *i.e.*, setting the instrument to zero, the indicator drifts from its zero setting and has to be restored by rebalancing. Drift is independent of a signal voltage and this implies, of course, that something tantamount to a signal is, somewhere, introduced into the circuit. Actually drift is caused by variation of some characteristic or "constant" and this may be associated with the valve, the external circuit or both. Thus, referring to Fig. 12, it is apparent that variation of the values of any of the circuit resistances will disturb the balance. Such variations may occur owing to temperature changes and because of this all resistors should be mounted away from any component that generates heat. The cathode resistors in (*d*) should be mounted close together so that, should an ambient temperature change in their vicinity be unavoidable, they will be equally affected. Also, if possible, they should be wire-wound with wire of negligible temperature coefficient. If carbon resistors must be used they should be of the cracked type and underrated.* Evidently all components should be protected from draughts.

More important than the foregoing causes of drift are the effects of variation in the cathode temperature and/or emission. These effects are somewhat complex and are not readily separable.

* *Carbon Resistors*, R. H. W. Burkett, *Wireless World*.

Reference to Fig. 16 shows the effect on the anode current of an EF 37 valve, connected as a triode, consequent upon increasing the heater voltage from 6·3 to 7·0 volts. It will be noted that the anode current and g_m, *i.e.*, μ/ρ, have increased. These effects could be regarded as due to an increase in the electron initial velocities consequent upon an increase in temperature or, alternatively, to an increase in the effective anode voltage. They may also be regarded as due to a variation in the cathode-grid contact potential caused by variation of cathode temperature. Consideration of (13) shows that a change in heater supply voltage will, owing to its modification of g_m, cause drift in (*b*), Fig. 12. An analysis of (*a*) shows that drift also occurs with this circuit. However, for (*d*), it is apparent from (18) that drift does not occur. Drift is also absent with (*c*). These last two statements are, of course, true only if both valves have identical characteristics. Since the characteristics are likely to differ some drift usually occurs but its magnitude is smaller than that associated with the circuits of (*a*) and (*b*). It may be minimized by use of a double triode, and, in the case of (*c*), by a valve having a common cathode, such as the ECC91. Apart from drift, it will be noted from (15) and (19) that the change in μ/ρ tends to modify the sensitivity. However, this is relatively unimportant if μ/ρ is negligibly small.

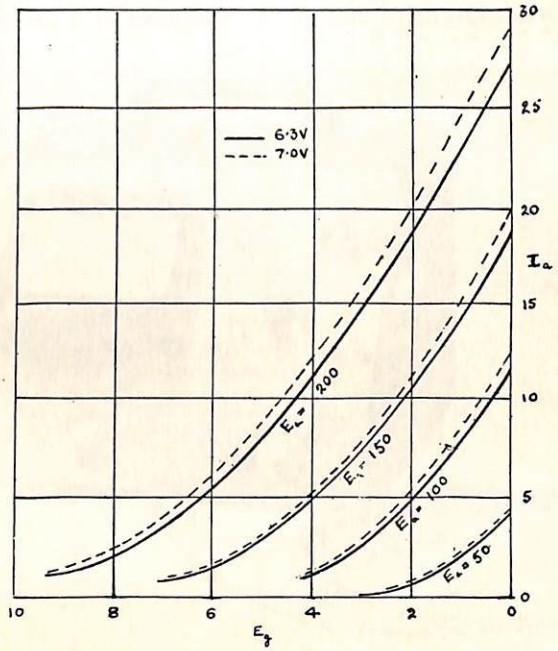


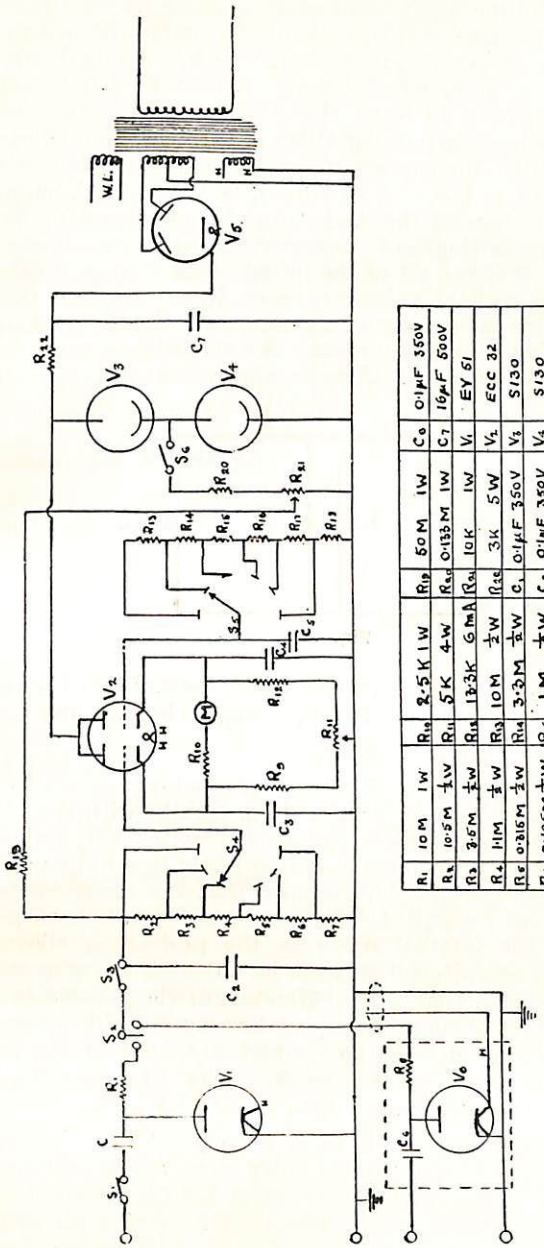
Fig. 16.

If the effect of a change in heater voltage is regarded as modifying the cathode-grid potential difference, then this is tantamount to the introduction of a signal voltage between cathode and grid of magnitude E_h, where E_h represents the change in cathode-grid potential difference consequent upon a change in heater voltage or random emission. Hence, in place of μE_s in (14) there must be written $\mu (E_s + E_h)$ and, whether E_s is equal to zero or otherwise, variations in E_h will cause drift. For circuit (*d*), the term μE_s in (16) becomes $\mu (E_s + E_h)$ while to the numerator of (17) a term μE_h must be added. Thus with this circuit drift does not occur provided E_h is the same for both valves. As previously indicated, this condition is more likely to be realized if a double triode is employed or, in the case of (*c*), a valve with a common cathode.

It will be noted that E_h may be regarded as due to random emission as well as to changes in heater voltage. This is because the emission current is subject to statistical fluctuations even when the heater voltage remains constant. Thus some drift may be experienced even should the heater voltage be stabilized.

Figs. 17 and 18 give details of a voltmeter, designed by the author, that makes use of a diode rectifier followed by a totally degenerative d.c. amplifier of the type shown by Fig. 12, (*d*). It will be noted that the diode is of the shunt type and that two are employed, one internally and one in the form of a probe. With

Fig. 17.



R_1	10M 1W	R_{12}	2·5K 1W	R_{19}	50M 1W	C_0	0·1μF 350V		
R_2	10·5M ½W	R_{11}	5K 4W	R_{20}	0·133M 1W	C_7	16μF 500V		
R_3	3·5M ½W	R_{14}	18·3K 6mA	R_{21}	10K 1W	V_1	EY 51		
R_4	1·1M ½W	R_{13}	10M ½W	R_{22}	3K 5W	V_2	ECC 32		
R_5	0·816M ½W	R_{14}	3·2M ½W	C_1	0·1μF 350V	V_3	S130		
R_6	0·105M ½W	R_{15}	1M ½W	C_2	0·1μF 350V	V_4	S130		
R_7	0·044M ½W	R_{16}	0·33M ½W	C_3	0·1μF 350 V	V_5	EZ 35		
R_8	10M ½W	R_{17}	0·1M ½W	C_4	0·1μF 350V	V_6	EY 51		
R_9	10K 6mA	R_{18}	0·047M ½W	C_5	0·1μF 350V	M	0-100 μAMP		

Fig. 18.

the internal diode the input capacitance is much higher than with the probe and, because of this, without the probe, the instrument is limited to an upper frequency of 20 Kc/s. Beyond this the instrument reads progressively low. Relatively large cathode resistors are employed in order that the calibration may, as far as possible, be independent of variations of the valve characteristics. In order to stabilize the characteristics the double triode was aged for 100 hours before use. A stabilized, 240 volt, H.T. supply is used but stabilization of the heater supply was found to be unnecessary. Zero setting and correction for drift are effected by means of R_{11}. Backing off of the initial diode voltage is effected by means of R_{19} and the voltage drop on R_{21}. Once set there is no need to disturb the setting of R_{21} unless the diodes are changed. Hence this setting is made by means of a screwdriver through one of the case ventilation holes. The backing off circuit is reproduced

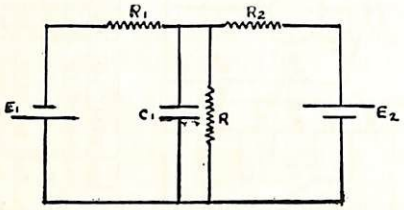


Fig. 19.

in Fig. 19 where E_1 and E_2 are, respectively, the e.m.f.'s of the diode and R_{21}, R is the attenuator, and $R_2 = R_{19}$. For balance

$$E_1/E_2 = R_1/R_2$$

and since $E_1 \fallingdotseq 1$ volt $E_2 \fallingdotseq 5$ volts. It is advantageous to have R_2 high because this resistor is virtually in parallel with R and thus reduces the input impedance. The ganged switches S_3 and S_4 are provided so that the instrument may first be set to zero by means of R_{11} before adjusting R_{21}. This ensures that the diode voltage is exactly backed off by R_{21}. S_2 is a changeover switch to permit changing from the internal diode to the probe. S_1 allows the signal voltage to be switched off in order to check the zero setting. It may be mentioned that with voltmeters of the present type an alternative method of neutralizing contact potential is to employ a double diode and impose the contact potential of the second diode on the grid of the second triode. This, of course, tends to neutralize the effect of the first diode.

The attenuator is arranged to give constant input resistance and it will be noted that similar resistances are simultaneously switched in the grid circuit of the second triode section. The anode currents are approximately 0·5*m*A each and thus the cathode-grid potential differences are about 6 volts. However, for this

voltage a positive-ion grid current flows in the attenuator resistors thus modifying the net c.g. potential difference. When the range of the meter is changed, this potential difference is altered owing to the effect of grid current. This unbalances the meter unless similar resistors are switched in both grid circuits. The switches S_4 and S_5 are ganged and, in practice, when the range is changed the change in the zero setting is inappreciable.

Probe.

The construction of the probe is shown by Fig. 9 and is almost self-explanatory. The capacitor is small, by virtue of being of the metallized paper type, and the 10M resistor is let into a slot in the Tufnol tube. The diode is a light interference fit in the tube and one of its heater leads is fed *via* the cable screening. Thus only a two-core cable is needed. The probe is readily detached from the instrument by means of a two-pin plug and socket.

Calibration and Accuracy.

Calibration was effected at 50 c/s. in conjunction with a dynamometer substandard. Accuracy is within 2% of the indicated values on all ranges and, with the probe, is unaffected by frequency up to 1·5 Mc/s., which was the highest available testing frequency. The ranges are 0·75, 2·5, 7·5, 25, 75, 250 volts, the smallest calibration being 0·1 volt. The three highest ranges are linear and employ a single scale. The three lowest ranges become non-linear as zero is approached and each has a separate scale. Non-linearity is most marked on the lowest range.

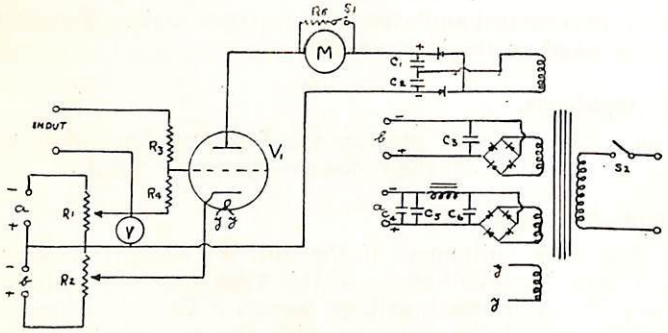
Input Impedance.

From (21) the input resistance is 5·15 megohms, and, as previously stated on p. 25, the probe capacitance is 5 $\mu\mu$F.

The Slide-Back Voltmeter.

A slide-back voltmeter, of the author's design, is shown by Figs. 20 and 21. Voltmeters of this type employ a null method and require an internal voltage source. The operation of the instrument may be appreciated with the assistance of Fig. 21. The potentiometer governing the voltmeter V is first adjusted so that the reading of V is zero. The bias potentiometer, R_2, is then adjusted until the valve anode current has some small reference value or is just cut off. R_1 is now adjusted so that V shows a large value and the signal then applied. The voltage on V is now slowly decreased until the meter just commences to indicate or reads the previous reference value. Under these circumstances the peak value of the signal voltage is approximately equal to the value shown by V.

Fig. 20.



R1 10K 30W	C1 4μF 350V	C6 4μF 350V
R2 250K 1W	C2 4μF 350V	V1 354V
R3 20M 10W	C3 1μF 300V	M 0-100μA
R4 400K 1W	C4 4μF 350V	V 0-400V
R5 200 w	C5 4μF 350V	

Fig. 21.

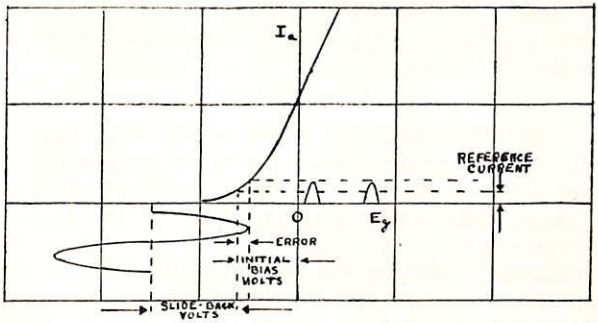


Fig. 22.

The slide-back voltmeter is not very suitable for measurements below about 10 volts because, at low values, considerable errors may occur. This may be made clear with the assistance of Fig. 22. From this it will be seen that, if a small reference current is employed, the peak value of the signal voltage must exceed the slide-back voltage in order that the mean value of the anode current pulses thereby produced may equal this current. The smaller the reference current the smaller is the error, the latter being, theoretically, zero when the reference current is zero. However, owing to the exponential character of the anode current-grid voltage curve, it is difficult to decide when the current is actually zero. Evidently the more sensitive the anode-current meter the smaller the error is likely to be, because a smaller reference current may be employed. For measurements of the order of a volt an error of 50% may easily occur and for such measurements a more suitable type of voltmeter should be employed. Errors are negligible with the voltmeter of Fig. 20 because this was designed to measure voltages up to 20-kV. by means of the attenuator R_3R_4. At 20-kV. the voltage applied to V is 400 volts and assuming the lowest value which can be conveniently read is 2-kV., then at this voltage V is 40 volts. Stabilized supplies are not employed because, with a voltmeter of the present range, their absence does not cause appreciable errors.

R.M.S. Meters.

Although the foregoing types of meters are frequently calibrated as reading r.m.s. values, these values are indicated only when the input waveform is sinusoidal. Since the meter deflection is proportional to the peak value of the input voltage, it is apparent that considerable errors may occur when harmonics are present. For example, if a voltage given by $V_1 \sin pt - V_3 \sin 3pt$ is applied to this type of instrument the peak value of the voltage is $(V_1 + V_3)$

volts and the meter indicates $(V_1+V_3) / \sqrt{2}$ volts. The true r.m.s. value is $\sqrt{V_1^2+V_3^2} / \sqrt{2}$ volts and the ratio of indicated to true value is, therefore,

$$(V_1+V_3) / (V_1^2+V_3^2)^{\frac{1}{2}}$$

Thus for a 10% harmonic the meter would read approximately 10% high. It is apparent that the error is dependent on the magnitudes and phases of the harmonics and, in the worst case, the percentage meter error may be equal to the percentage of harmonics present.

By operating a valve on the portion of the I_a/E_g characteristic that approximates to a quadratic law it is possible to construct an instrument that reads true r.m.s. values. If the relation between anode current and grid voltage is given by

$$i_a = a_0 + a_1 v_g + a_2 v_g^2 \qquad (22)$$

then with a signal voltage of $V_m \sin pt$

$$i_a = a_0 + a_1 V_m \sin pt + a_2 V_m^2 \sin pt$$

and the mean value of i_a over a cycle is

$$I_a = a_0 + a_2 V_m^2/2 \qquad (23)$$

or, if a_0 is balanced out,

$$I_a \propto V_m^2$$

Thus the instrument may be calibrated to indicate r.m.s. values and, provided the magnitude of the signal voltage is such that the valve always operates on that part of the characteristic given by (22), the readings will be independent of waveform.

A typical circuit, employed by the author, is shown by Fig. 23. The purpose of C_1 is, of course, to block any d.c. component that may be present in the input voltage. In order that the input impedance shall be high, it is apparent that the peak value of the input voltage must not exceed the value of the grid bias voltage. For the circuit shown, the initial value of the bias is 3·5 volts, but this is increased by a signal voltage owing to the increase in the mean value of the anode current given by (23). Hence a voltage slightly higher than 2·5 volts r.m.s. may be applied before grid

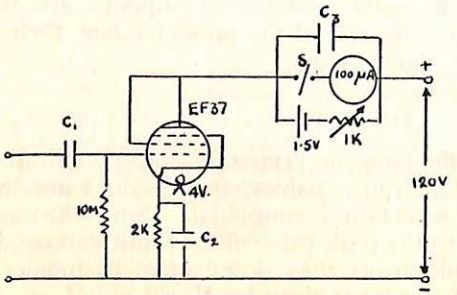


Fig. 23.

current flows. It will be noted that $I_a \propto a_2$. Hence for high sensitivity it is necessary that a_2, or d^2i_a/dv_g^2, should be large.

The functions of C_2 and C_3 are to act as by-passes for the alternating components of the anode current. Writing $\sin^2 pt$ as $(1 - \cos 2pt)/2$

$$i_a = a_0 + a_1 V_m \sin pt + a_2 V_m^2 (1 - \cos 2pt)/2$$

and the meter current will contain components of signal and twice signal voltage frequency. At low frequencies and with a light meter movement, vibration of the meter pointer may be perceptible unless C_3 is employed. The value of C_3 depends on practical conditions. Thus with a 50 c/s signal the author found a capacitance of 32μF. (electrolytic) necessary to render vibrations imperceptible. The calibration characteristic is shown by Fig. 24.

The method of zero setting and the purpose of S are explained on p. 46.

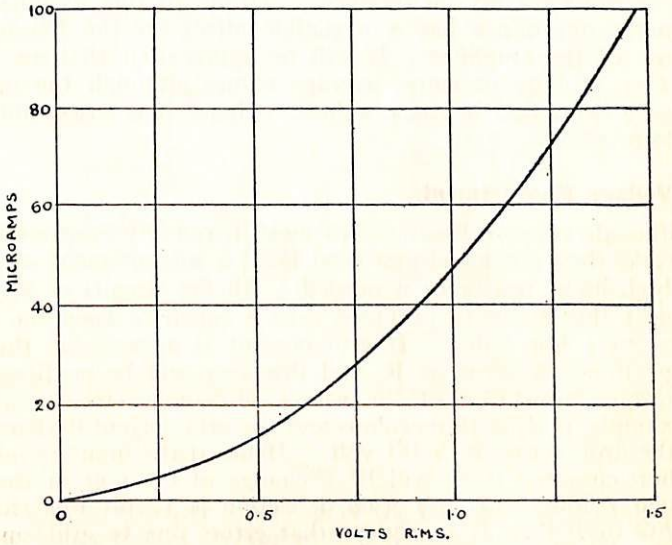


Fig. 24.

Since the first stage of a rectifier-amplifier voltmeter is a diode, and diode voltmeters, normally, are unsuitable for reading voltages lower than 0·1 volt, it follows that rectifier-amplifier voltmeters will not read voltages less than 0·1 volt. For this reason such voltmeters rarely have the instrument calibrated for less than 0-1 volt F.S.D. For reading low voltages, 1-1000 millivolts, an amplifier-rectifier voltmeter is usually employed. As the name implies, this voltmeter comprises an amplifier followed by a rectifier. Compared with the R-A type the A-R voltmeter has a

more restricted frequency range owing to the difficulty of producing a wide-band amplifier. Hence the upper frequency limit is usually about 5 Mc/s.

The first stage of an A-R voltmeter is usually a cathode follower in the form of a probe. The output impedance of this has already been shown to be $\rho/(\mu+1)$ ohms and, since this is relatively low, the shunting effect of the screened cable thereon is small. For example, the output resistance of an EF50, connected as a triode, is less than 100 ohms. If the probe cable has a capacitance of, say, 25$\mu\mu$F, for a 2 ft. length the reactance of this is 1300 ohms at 5 Mc/s which, shunted across 100 ohms, would have a negligible effect. The output of the cathode follower is fed to an attenuator and from this to the first stage of a wide-band amplifier. The amplifier usually terminates with a cathode follower, and across the output of this is a moving coil meter fed through either a half-wave or full-wave crystal rectifier. Owing to this termination the meter impedance has a negligible effect on the frequency response of the amplifier. It will be appreciated that an A-R voltmeter actually measures average values although the meter is usually calibrated in r.m.s. values. Hence it is susceptible to waveform errors.

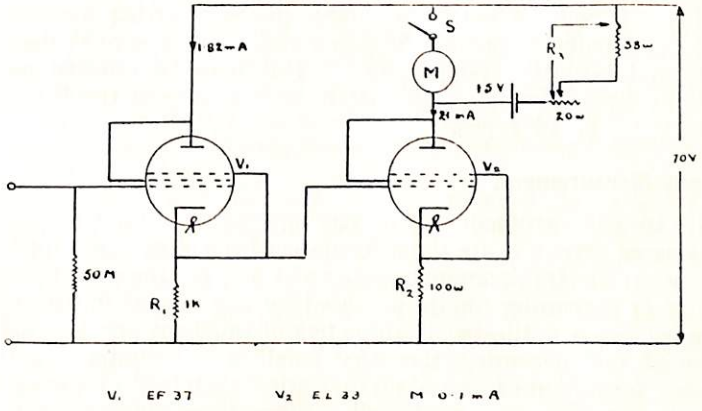
D.C. Voltage Measurement.

Although valve voltmeters are most frequently employed for a.c. work, they are sometimes used for d.c. measurement when a very high input resistance is needed. All the circuits of Fig. 12 will meet this necessity provided care is taken to keep the grid current to a low value. If grid current is appreciable then a voltage drop will occur on R_g and this drop will be modified by the internal resistance of the source undergoing measurement. For example, if R_g is 10 megohms and the grid current 0·01 micro-amp the drop across R_g is 0·1 volt. Hence if the input terminals are short-circuited there will be a change of 0·1 volt in the instrument reading. If full scale deflection is 1 volt this change amounts to 10%. It is evident that errors due to grid current depend on the source resistance and that satisfactory results are likely to be obtained only when there is no shift in the meter reading consequent upon short-circuiting R_g. Methods of obtaining low grid currents are described on p. 47, and here it may be mentioned that the d.c. voltmeter in the timer of Fig. 47 operates with H.T. and L.T. supplies of 28 and 4 volts respectively. Under these circumstances it was found possible to operate the valve with an open grid.

Range changing may be effected by means of either of the methods shown by Figs. 15 and 18. Provided the valves are worked on the linear portions of the characteristics, equal deflec-

tions occur when the grid terminal is made either positive or negative. However, unless a centre-zero meter is employed, a two-pole changeover switch is necessary to change the meter connections. As with a.c. voltmeters, it is advisable to employ stabilized supplies.

An example of a further type of d.c. voltmeter is shown by Fig. 25. Here it will be noted that two valves are employed, both connected as cathode followers. The first valve is intended to couple the signal to the second valve and, to avoid grid current effects, should be operated at reduced heater and anode voltages. The bias voltage of the second valve is the algebraic sum of the cathode resistor voltages, and conditions should be arranged so that this voltage is negative to cathode. Zero setting is effected by means of R_3 and the range may be altered by means of R_m. On the lowest range R_m is, of course, the meter resistance. It follows that even with full scale deflection the volt drop across the anode load is negligibly small.



V₁ EF 37 V₂ EL 33 M 0·1 mA

Fig. 25.

In order to consider the value of the current in R_m consequent upon a change in grid voltage and anode current in V_2, let I be the initial value of the anode current with $I_m = 0$. Then

$$I = E/R_3$$

If I becomes $I + \triangle I$ then

$$R_3 \triangle I_2 = R_m I_m$$

and $$\triangle I_2 = R_m I_m / R_3$$

Also $$I + \triangle I = E/R_3 + I_m + \triangle I_2$$

$$= E/R_3 + I_m + R_m I_m / R_3$$

or $$I + \triangle I - E/R_3 = I_m (1 + R_m/R_3)$$

or $$I_m = \frac{\triangle I}{1 + R_m/R_3}$$

where I_m and I_2 are, respectively, the currents in R_m and R_3 and E is the backing cell e.m.f., 1·5 volts for the circuit shown.

Hence in order that I_m should be a large fraction of $\triangle I$, R_m/R_3 should be small. R_3 is, of course, equal to E/I. From p. 30 the *gains* of the first and second valves are, respectively,

$$\mu_1 R_1/[\rho_1 + (\mu_1 + 1)\ R_1] \text{ and } \mu_2 R_2/\ [\rho_2\ +\ (\mu_2 + 1)\ R_2\ ]$$

Hence

$$I_m\ =\ \frac{\mu_1 \mu_2 R_1 E_s}{(1 + R_m/R_3)\ [\rho_1 + (\mu_1 + 1)\ R_1]\ [\rho_2 + (\mu_2 + 1) R_2]}$$

and on substituting the values given in Fig. 25

$$I_m\ =\ 1{\cdot}23\ m\text{A per volt.}$$

The values shown in Fig. 25 are those employed by the author and, with these, full scale deflection occurred for an input to V_1 of 1 volt. The scale is linear, independently of whether the input to the grid of V_1 is positive or negative. The circuit may be employed for a.c. measurements by preceding V_1 with a diode rectifier. It will be noted that with the zero-setting method of Fig. 25 the switch S must not be closed until the heater of V_2 reaches its normal operating temperature. This may be effected automatically by a thermal delay switch, such as any of the Ediswan DLS 10, 15, or 16 types.

Current Measurement.

Electronic instruments for measuring current are not employed to the same extent as are those for measuring voltage and, furthermore, what electronic ammeters do exist are, for the greater part, voltmeters measuring the drop caused by the current in a resistor of known value. Electronic ammeters of this type are principally employed for measuring the very small d.c. currents resulting from the formation of ions in an ionization chamber* or caused by weak illumination on a photo-cell. Since these currents may be of the order of 10^{-15} to 10^{-9} ampere it is apparent that an ordinary valve, operated at rated conditions, with a typical grid current of about 10^{-8} to 10^{-7} ampere is unsuitable. Referring to Fig. 4 it appears that if operation could be maintained at the point P satisfactory results would be obtained. However, it will be noted that dI_g/dE_g is relatively large in the vicinity of P and the volt drop on the grid resistor, caused by the current undergoing measurement, would shift the operating point and cause a relatively large grid current to flow. Because of this valves employed for low current measurements (termed electrometer valves) are biassed to the left of P so that no electrons are collected from the cathode and what current flows is a *negative* grid current. Some of the

* *Nuclear Particle and Radiation Detectors*, D. Taylor and J. Sharpe, J.I.E.E. Part II, April, 1951, p. 174.

components of grid current have already been quoted and here those contributing to negative grid current may be restated and amplified.

(1) Surface leakage, internal and external.

(2) Electron currents caused by
   (*a*) thermionic grid emission due to heating by the cathode heater.
   (*b*) photoelectrons emitted by the grid under the action of the cathode light and/or external light, and by soft X-rays due to the anode current.

(3) Positive ion currents caused by
   (*a*) collision processes with the residual gas molecules.
   (*b*) positive ion emission from the cathode.

It is now necessary to consider methods of reducing grid current when it is desired to use ordinary valves for electrometer purposes. Internal leakage is, of course, controlled by the valve manufacturer but is usually negligible. External leakage occurs across insulating surfaces such as moulded bases and the glass envelope. Where such bases exist they should be removed or an all-glass type of valve employed. Leakage over the glass surface may be reduced by suitable cleaning and drying followed by treating the surface with some non-hygroscopic substance such as paraffin wax, ceresin, moisture resistant varnish, or, preferably, a silicone* coating such as B.T.H. "Teddol."

Thermionic grid emission can be reduced by reducing the heater current. Evidently this is also likely to reduce photo-electrons due to light from the cathode surface. Photo-electrons due to the action of external light can be eliminated by operating the valve inside a light-tight cover, while those due to X-rays, created by the anode current, may be reduced by reduction of the anode potential.

Probably the most important component of grid current is that due to positive ions. Positive ion emission from the cathode can be reduced by reducing the heater current, suitable results being obtained if a 6·3V heater is reduced to 3-4V. This has the effect of reducing the heater current to about 65-75% of its normal value. Ionization by collision depends on the anode-cathode potential difference, the gas pressure, and the total cathode emission. Here, again, a reduction in heater voltage is beneficial and, from foregoing statements, it is apparent that this reduction simultaneously confers several advantages. However, the degree to which the heater current can be reduced is evidently limited because sufficient emission must be available to operate the meter in circuits of the types shown by Fig. 12. Since the ionization potentials of gases run from about 10 volts upwards it appears that ionization

* *Silicones*, *Wireless World*, September, 1945, p. 265.

by collision can be prevented provided the anode-cathode p.d. is not greater than about 10 volts. Reducing the normal anode voltage to a figure larger than 10 will result in an appreciable reduction in grid current, but for very low currents, 10 volts or less should be aimed at. However, a lower limit is set for the anode voltage by the possibility of cut-off of the anode current. Since $I_a \propto (E_a - \mu E_g)^{3/2}$, it is apparent that E_a must be larger than μE_g. Because the electrons are emitted from the cathode with energies of at least 1 volt, E_g must be at least 1 volt negative and, generally, E_a must be greater than μ volts. Hence for electrometer work valves with low values of μ are required in order that the anode voltage may be reduced.

The foregoing argument concerning the effect of anode potential is mainly applicable to triodes. However, pentodes may also be employed for electrometer work. With these valves variation of anode potential has, of course, a relatively small effect on anode current and the relation $I_a \propto (E_a - \mu E_g)^{3/2}$ is not applicable. However, a similar relation holds between the screen and control grid voltages and we have $E_s/E_g \fallingdotseq K$ where K is a constant. The anode should be maintained at a potential no greater than that of the screen; preferably, less.

Grid Current Measurement.

A method of measuring grid current, employed by the author, is shown by Fig. 26. To operate this circuit, R_2 is reduced to zero, S_1 closed, the bias set to a value at which observation is desired and the bridge circuit then balanced by means of R_1. Under these conditions grid current is flowing in the high resistance R_g and the

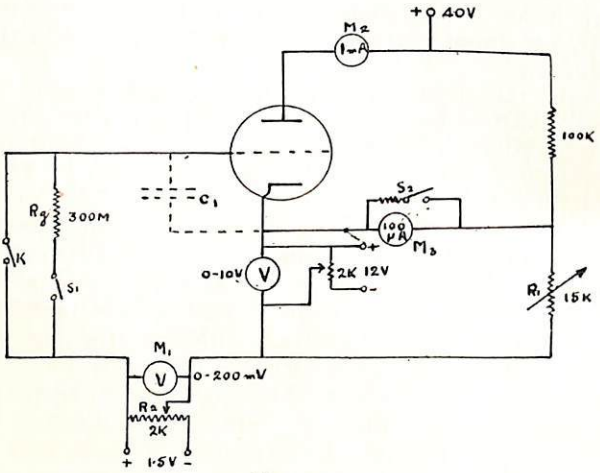


Fig. 26.

volt drop on R_g either opposes or assists the bias voltage depending on whether grid current is negative or positive. The key K is then closed, short-circuiting R_g and thus unbalancing the bridge. Balance is then approximately restored by means of R_2 which introduces a voltage into the grid circuit equal to that previously existing across R_g. Final balance is obtained by opening S_2 and readjusting R_2, if necessary. If E is the voltage introduced by R_2, then

$$R_g I_g = E$$
$$\text{and} \quad I_g = E/R_g$$

Fig. 27 shows an I_g—E_g characteristic obtained in this manner by the author from a 1L4 valve, triode-connected, operated with full heater volts and an anode potential of 40 volts. The current is relatively large but could be reduced by employing the expedients previously described.

The foregoing method of measurement is suitable for currents down to about 10^{-13} to 10^{-12} ampere. For smaller currents it lacks

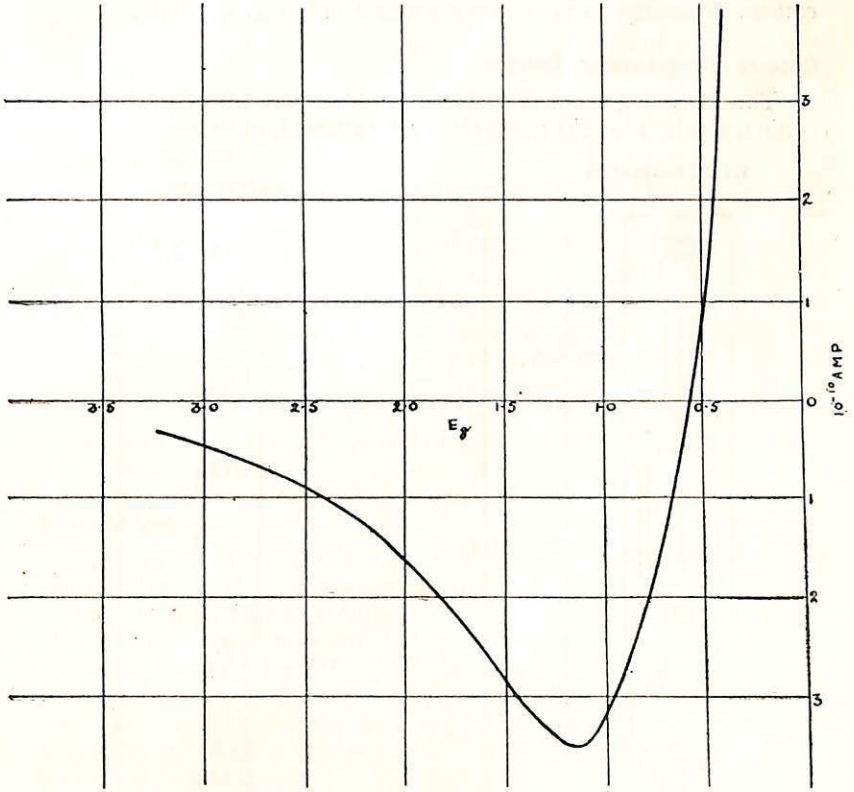


Fig. 27.

sensitivity and a different method must be employed. The circuit of Fig. 26 is still used, but with R_g open-circuited, and the method of measurement is to observe the rate of change of grid potential with the grid floating. Firstly S_1 is opened, K closed, the circuit balanced by means of R_1, and, with the bias set to the desired observation point, an mV$-\mu$A characteristic taken by means of R_2, M_1 and M_3. The readings of M_1 and M_3 are now set to zero and K is opened. On opening K, grid current commences to charge the grid-cathode capacitance, C_1, and, in consequence, the potential of the grid changes, this being indicated by the deflection of M_3. A small change in grid potential is permitted over a measured time interval $\triangle t$. If $\triangle$V is this potential change, as determined from the mV$-\mu$A characteristic previously taken, then

$$I_g = C_1 dV/dt \doteqdot C_1 \triangle V/\triangle t$$

provided $\triangle$V is small. In measurements of this kind it is, of course, necessary to take every precaution to avoid drift.

Special Electrometer Triodes.

The foregoing discussion has been directed towards the use of ordinary valves for electrometer work rather than to valves specially

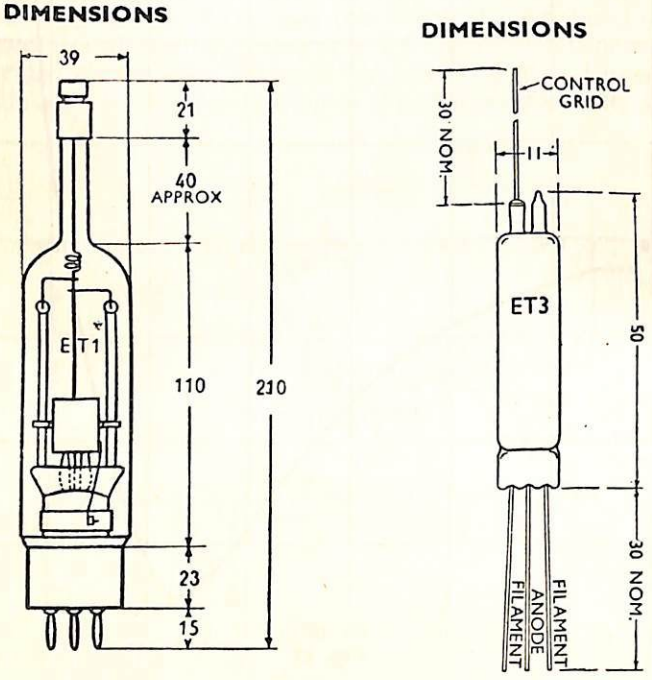


Fig. 28.

designed for this purpose. The demand for such valves is not high and, at present, few types are available. Two available types are the ET1 and ET3 and diagrams of these with their dimensions in mms. are shown by Fig. 28, the latter type being a sub-miniature. The ratings are as follows :—

	ET1		ET3
Filament Voltage	1·0 volt		1·25 volts.
Filament Current	0·1 amp.		0·025 amp.
Anode Voltage	4	6	9 volts.
Mutual Conductance	0·05mA/V	0·1mA/V	70μA/V
Amplification Factor	—		1
Grid Resistance	—		$> 10^{14}$ ohms.

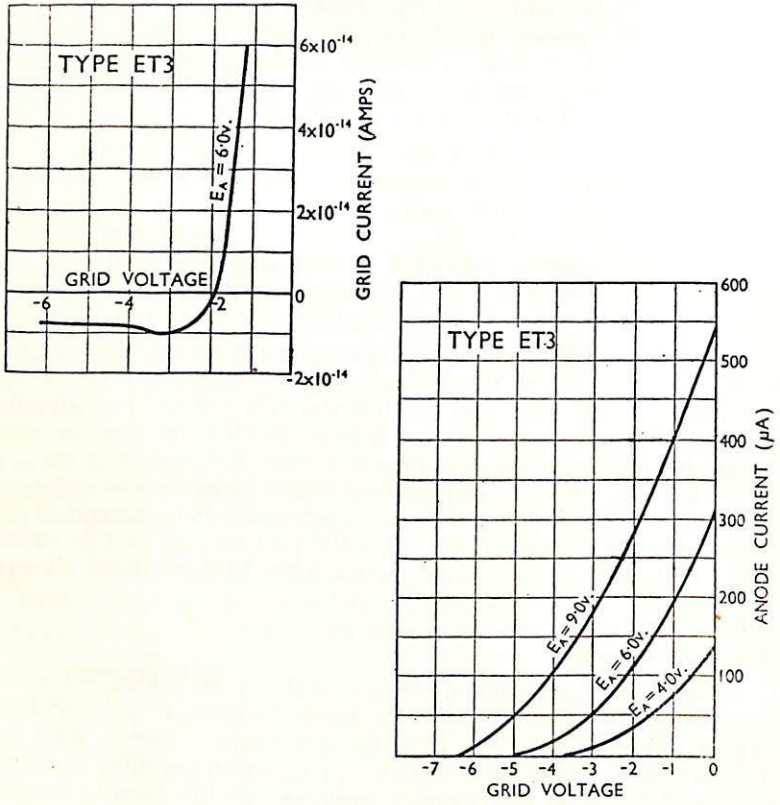
It will be noted that these ratings confirm statements previously made in the general discussion. The recommended operating conditions of the ET1 include that the grid potential should be greater than two volts negative with respect to the negative end of the filament, the glass must be clean and dry, operation should be in the dark, screening employed, and high-capacity accumulators employed for the supplies in order to reduce fluctuations and drift. With an anode potential of 3 volts the grid current of this valve is less than 10^{-16} amp. Some characteristics of the ET3 are shown by Fig. 29.

Ammeter Circuits.

Since most electronic ammeters operate by measuring the volt drop caused by the current on a known resistor it follows that ammeters of this type are virtually voltmeters. Hence, used in this way, most of the voltmeters previously described may be used as ammeters. For electrometer work any of the circuits shown by Figs. 12 and 25 may be employed provided the methods previously discussed to keep grid current small are adopted. Fig. 30 shows a typical circuit of a portable radiation monitor. It will be noted that the signal is derived from ionization currents flowing in R_1 and that the filament battery also serves for backing-off purposes.

Wattmeters.

Electronic wattmeters take two forms, one measuring the power dissipated in a known resistance and the other the power flowing along a pair of conductors. The first form merely consists of a valve voltmeter connected across a resistor and is calibrated according to the law $W = V^2/R$. This type of instrument is sometimes termed an "output" meter, because it is frequently employed to measure the power output from an amplifier. The second form is more complex and is somewhat similar to a double-triode valve voltmeter. Its outstanding properties are high



CHARACTERISTIC CURVES OF AVERAGE VALVE.

By courtesy of G.E.C., Ltd

Fig. 29.

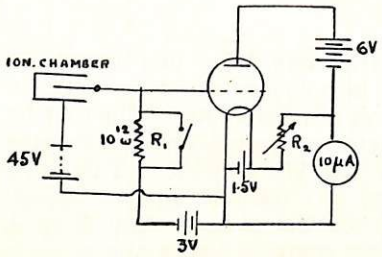


Fig. 30.

sensitivity and input impedance and a fairly wide frequency range. Thus an instrument may be readily constructed with a full scale deflection of 100 milliwatts and an input impedance of the order of a megohm. Furthermore the scale is linear.

Reference to Fig. 31 shows the circuit of a wattmeter, of the author's design, for measuring powers from 2 milliwatts up to 50 watts. A double triode is employed and was selected so that the characteristics of the two valves are as similar as possible. This is important because difficulties in obtaining satisfactory operation are largely due to dissimilarities of characteristics. R_1R_1 are identical, non-inductive resistors, and R_4 is a non-inductive potential divider. The load current passes through R_1R_1 and the pressure is applied across R_4. The value of R_4 is made sufficiently high so that its current consumption is negligible compared with the load current passing through R_1. Thus, if the instantaneous value of the load current is i, the volt drop across each of R_1R_1 is iR_1. The voltage across the junction of R_1R_1 and R_5 is vR_5/R_4* where v is the instantaneous value of the supply voltage. Hence the grid voltages (apart from the biassing voltage E_g) of the two valves, V_1', and V_1'', are, respectively,

$$(vR_5/R_4 + R_1i) \text{ and } (vR_5/R_4 - R_1i)$$

Thus V_1' experiences the sum of the voltages and V_1'' the difference. When $i = 0$ these voltages are, of course, identical.

The value of E_g is such that the valves work on the parabolic portions of their characteristics. The individual anode currents are then given by an expression of the form

$$i_a = a' [E_a + \mu (V_g + E_g)]^2$$

if the values of R_9R_{11} are such that E_a, *i.e.*, the anode voltage, may be regarded as constant. This expression may be simplified and written as

$$i = a (b + v_g)^2$$

Now assuming that the individual valves have different characteristics, their respective anode currents may be written as

$$i_1 = a_1 (b_1 + v_g)^2 = a_1 (b_1 + vR_5/R_4 + R_1i)^2$$
$$i_2 = a_2 (b_2 + v_g)^2 = a_2 (b_2 + vR_5/R_4 - R_1i)^2$$

and the meter current is

$$\frac{R_9i_1 - R_{11}i_2}{R_9 + R_{11} + R_m}$$

Substituting for i_1 and i_2 the numerator of this expression is

$$\begin{aligned} &a_1b_1^2R_9 - a_2b_2^2R_{11} + 2 (R_9a_1b_1 - R_{11}a_2b_2) R_5/R_4.v \\ &+ 2 (R_9a_1b_1 + R_{11}a_2b_2) R_1i + (R_9a_1 - R_{11}a_2) R_5^2/R_4^2.v^2 \\ &+ (R_9a_1 - R_{11}a_2) R_1^2i^2 + 2 (R_9a_1 + R_{11}a_2) R_1R_5/R_4.vi \end{aligned} \qquad (23)$$

and if v and i are purely alternating quantities then, since the meter reads the mean value of the current, components correspond-

* Assuming R_5 to be in operation.

ing to the second and third terms do not appear in the output. The fourth and fifth terms are proportional to the losses in R_4 and R_1R_1, while the last term is proportional to the power in the load. Evidently when the valves have identical characteristics the mean value of the meter current is

$$I_m = \frac{2aR_1R_5R_{11}}{(R_9+R_{11}+R_m)\,R_4}\ \overline{vi}$$

where $\overline{vi}$ is the mean power in the load and $R_9 = R_{11}$.

It is apparent that it is particularly desirable that the valve characteristics should be identical because otherwise allowance must be made for the losses in R_1R_1 and R_4 and, in addition, a zero error, proportional to the first term of (23) exists. This error may be eliminated by adjustment of the tapping point of R_9R_{11} and, by making R_1 and R_5/R_4 small, it is possible, except for small powers, to make the fourth and fifth terms negligible compared with the last. Alternatively the tapping point may be adjusted so that $R_9a_1 = R_{11}a_2 = K$. Then

$$I_m = \frac{K\,(b_1^2-b_2^2)}{R_9+R_{11}+R_m} + \frac{4K}{R_9+R_{11}+R_m} \cdot \frac{R_1R_5}{R_4}\ \overline{vi}$$

and the first term of this may be eliminated by adjustment of the meter zero adjuster.

In the author's instrument difference in the characteristics is minimized by valve selection and adjusting E_g until a suitable operating point is found at which the difference is as small as possible. The efficacy of these methods may be checked by applying a voltage to the wattmeter without a current. There should, of course, be no deflection of the meter. Also a current may be passed through R_1R_1 without the application of a voltage to R_4. Again no deflection should occur. The phase-sensitive properties of the meter may be checked by means of a purely capacitive load and also by means of a load consisting of a capacitor in series with a non-inductive resistor. In the first instance no deflection should occur while in the second the watts recorded should be equal to V^2/R, where V and R are, respectively, the resistor voltage and resistance.

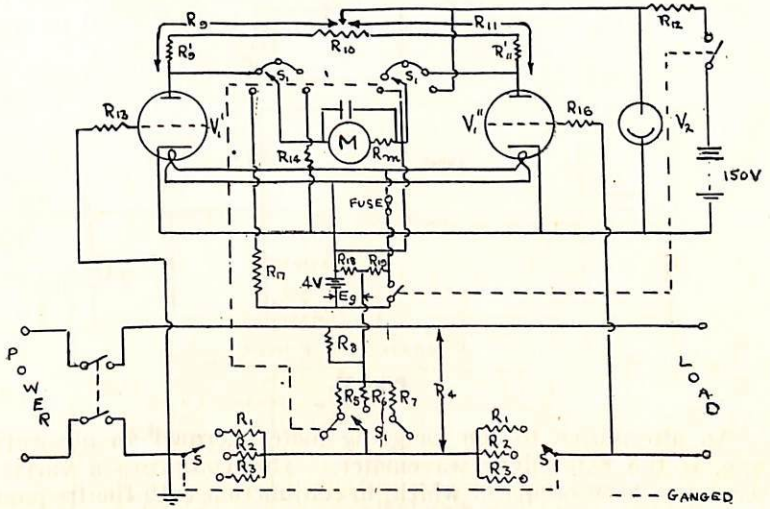
Values of R_1R_1 and R_8.

These values should be chosen to limit the grid swing to a fairly small amount because otherwise the valves may be operated on parts of the characteristics that do not obey a quadratic law. Also if R_8 is too high stray capacitance may shift the potential of the tapping point on R_5R_8.

It will be noted that the wattmeter of Fig. 31 is operated from batteries. If attempts are made to operate the instrument from

the mains supply considerable difficulties may be experienced. These difficulties arise when the circuit upon which measurements are made is earthed, or has capacity to earth. With mains operation the tapping point of R_4 is connected *via* R_{18} to the centre-tap of the secondary of the mains transformer and thus has capacity to earth and the primary winding. In a similar manner the heaters and cathodes have capacity to earth. Hence when the circuit upon which measurements are made also has capacity to earth, or is earthed, the potential distribution across the two parts of R_4 is altered and erroneous readings are obtained. This is readily demonstrated by reversing the connections to the supply terminals of Fig. 31. Invariably, when this is done a different reading is obtained from that before reversal.

The instrument of Fig. 31 has three current ranges 2, 20 and 200 *m*A, and three voltage ranges 10, 50, and 250 volts. The switches SS are ganged and the current range is selected by means of these. S_1, of course, selects the voltage range. The function of C is to by-pass the alternating components in the meter circuit current. In order to check the voltages of the L.T. and H.T. supplies, the meter, M, may be converted to either a 5 or 100 volt-meter by means of the switches S_1.



R_1	100 ω	R'_{11}	10 K	V_1	E C C 32
R_2	10 ω	R_{12}	10 K	V_2	85 A2
R_3	1 ω	R_{13}	10 K	M	0-100 μA
R_5	3 K	R_{14}	1 M	C	50 μF
R_6	15 K	R_m	610 ω		
R_7	80 K	R_{16}	10 K		
R_8	1 M	R_{17}	4·27 K		
R'_9	10 K	R_{18}	315 ω		
R_{10}	2·2 K	R_{19}	470 ω		

Fig. 31.

Frequency Measurement.

The means employed to measure an unknown frequency depend on the value of that frequency since, generally, it is not possible to employ a single instrument for the entire frequency spectrum. As is well-known, for radio frequencies it is usual to employ some form of wavemeter. This instrument may consist of an LC circuit which is tuned to resonate at the frequency undergoing measurement. Energy is absorbed from the frequency-producing circuit either by direct or mutually-inductive coupling. To indicate resonance a crystal diode and d.c. microammeter may be employed in the manner shown by Fig. 32. Alternatively some form of valve voltmeter may be used as a resonance indicator. Referring to Fig. 32, if the secondary-coil metering circuit is employed, then, of course, the capacitance potential divider metering circuit is unnecessary, and *vice-versa*. The principles of the latter circuit have already been given on p. 16. The circuit shown by Fig. 32 is that of a wavemeter employed by the author for frequency measurements from 200 Kc/s to 2 Mc/s. The frequency is determined from the values of L and C (one or both of which is continuously variable) and the formula $f = 1/2\pi\sqrt{LC}$.

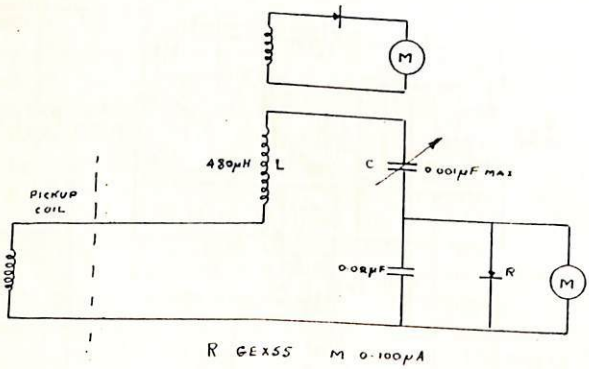


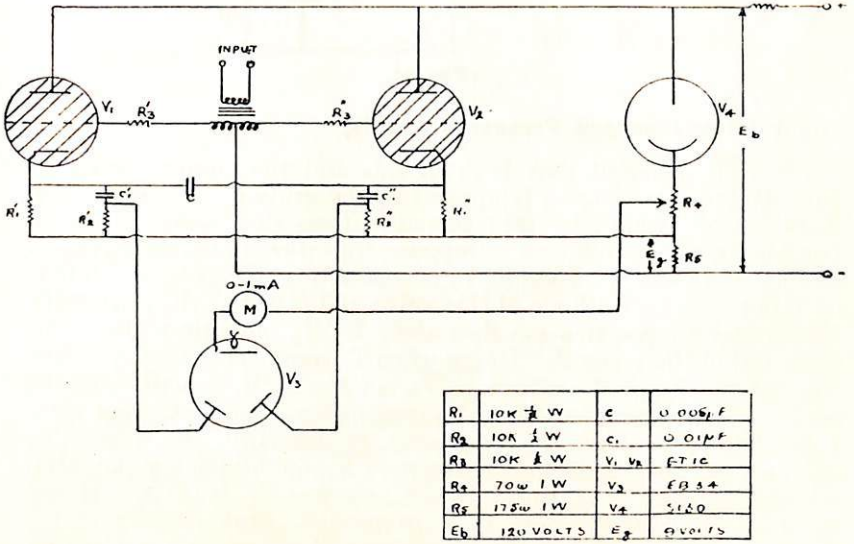
Fig. 32.

An alternative to the foregoing meter, termed an absorption type, is the heterodyne wavemeter. This comprises a variable-frequency, local oscillator which, in conjunction with the frequency to be determined and a mixing device, produces a beat note. This note may be detected by means of either telephones, loud-speaker, magic eye or meter and, by adjustment of the local oscillator, either reduced to zero frequency or to some known value. The beat note frequency is $f = (f_o - f_s)$ and hence if f and f_o are known f_s is determined. f_o and f_s are, respectively, the local oscillator and signal frequencies.

For frequencies in the audio and low r.f. range (<100 Kc/s) a meter may be employed in which the frequency is read directly on a moving coil instrument. Apart from this advantage meters of this type are largely independent of waveform and magnitude of input voltage and have a fairly high input impedance. Several of these instruments are described below.

Thyratron Frequency Meter.

A frequency meter employing thyratrons is shown by Fig. 33 and is suitable for frequencies up to about 5 Kc/s. The two halves of the circuit are nominally identical and each tube fires in turn when the potential of its grid relatively to the transformer centre-tap is positive. Considering the circumstances when V_2 is conducting, the potentials across R_1'', C_1'', and C are then equal and are also equal to $(E_b - e_a)$ volts, where e_a is the tube arc drop. When the grid of V_1 becomes positive this tube fires and a voltage approximately equal to E_b appears across R_1'. Momentarily the voltage across R_1'' is then $2E_b$ and, since the voltage between anode and negative of V_2 is E_b, the voltage between anode and cathode is $E_b - 2E_b = -E_b$. Thus V_2 is extinguished, C and C_1'', discharge through R_1'', C_1'' continuing its discharge through R_2''. Half a cycle later V_2 fires, V_1 is extinguished and a similar discharge occurs in the left-hand circuit.



R_1	10K ½ W	C	0.005 µF
R_2	10K ½ W	C_1	0.01 µF
R_3	10K ½ W	V_1, V_2	ET 1C
R_4	70Ω 1 W	V_3	EB 34
R_5	175Ω 1 W	V_4	5R60
E_b	120 VOLTS	E_g	9 VOLTS

Fig. 33.

Meter Current.

At the moment a tube fires, the charging current for its associated capacitor passes through the diode and meter in preference to through the relatively high resistance of either R_2' or R_2''. Hence the capacitor is charged with great rapidity to a potential of $(E_b - e_a)$ volts. The quantity acquired by the capacitor is $C_1(E_b - e_a)$ coulombs and thus the average value of the meter current is

$$I = 2fC_1 (E_b - e_a) \text{ amperes}$$

where f is the frequency of the input signal. This result assumes, of course, that the capacitors are completely discharged every time a tube is cut off. It will be noted that I is proportional to f and C_1 and thus a linear scale results. The range of the instrument may be extended by simultaneously changing C_1' and C_2'' by means of ganged switches.

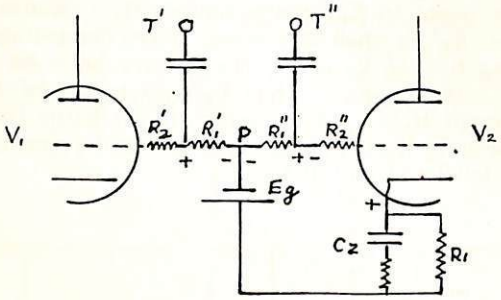


Fig. 34.

Input Impedance and Frequency Limits.

It will be noted that both in this and the circuit shown by Fig. 35 the input signal is applied to the grids via a transformer. This is necessitated by grid current effects that occur if an RC combination is employed. Suppose that the input arrangement of Fig. 34 is used. Then, if V_2 is conducting, the potential difference between the cathode of this valve and P is $E_g + (E_b - e_a)$ volts. This causes a positive ion flow along $R_1''R_2''$ and thus alters the potential of the point P. Hence when T' commences to go positive the effective signal voltage is $V_m \sin pt + R_1''I_i$, where I_i is the positive ion current. Hence V_1 fires earlier than if I_i were zero. Immediately V_1 fires, C_2 commences to discharge and a positive ion current flows in the grid resistors of V_1, modifying the potential across R_1'' from $-V_m \sin pt$ to $-V_m \sin pt + R_1' I_i$ volts. Hence V_2 may be restruck and V_1 extinguished, thus recharging C_2. Oscillations thus occur at the beginning and end of each conducting period and may persist until $V_m \sin pt$ rises sufficiently to prevent

their occurrence. Evidently C_1' and C'' in Fig. 33 are charged more frequently than once per ccyle and M reads high.

It is apparent that an input transformer imposes limitations on the circuits of Figs. 33 and 35, because the input impedance is a function of voltage and frequency. To minimize this a high primary inductance should be employed and, generally, a good quality intervalve, push-pull radio transformer will be found suitable, provided the secondary impedance is fairly low. If the last proviso is not observed self-oscillations may occur. High frequency limitation is due to the deionization time of the thyratrons. The instantaneous voltage applied to each grid is $(V_m \sin pt - E_g)$ and firing occurs when this quantity becomes equal to $-E_b/K$ where K is the thyratron control ratio. Hence the minimum operating voltage is $(E_g - E_b/K) / \sqrt{2}$ volts per valve or $\sqrt{2}\,(E_g - E_b/K)$ volts for the two valves. It is apparent that the instrument is independent of the input signal waveform. The function of R_4 is to neutralize the effect of the electron initial velocities which, otherwise, would cause a deflection of the meter in the absence of a signal. Employing the valves and components shown by Fig. 33, the author obtains full scale deflection for 500 c/s. If desired, square waves may be derived from R_1.

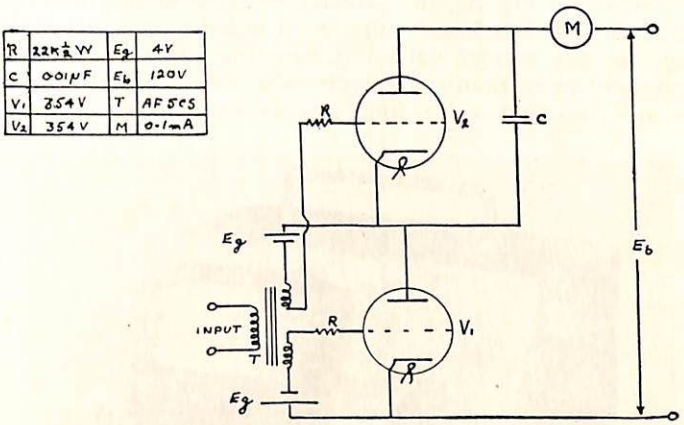


Fig. 35.

A further frequency meter, in which either hard or gas-filled valves may be used, is shown by Fig. 35. When V_1 commences to conduct the capacitor is charged through the meter M until the anode-cathode potential difference is too low to maintain the current. V_1 is then cut off with C charged to $(E_b - e_a)$ volts. During the next half-cycle the grid of V_2 approaches a positive potential whereupon this valve conducts, discharging C. It is

apparent that the meter current is given by

$$I = fC\ (E_b - e_a) \text{ amperes}$$

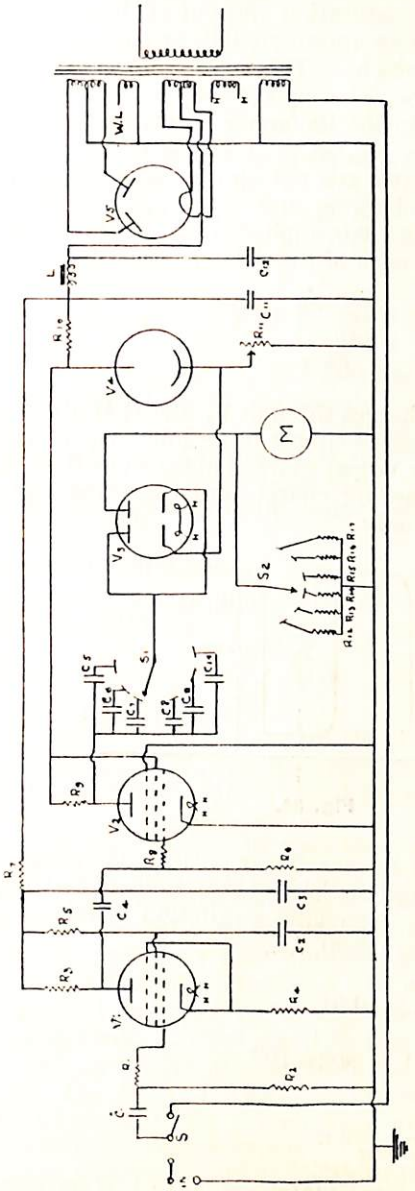
provided that the capacitor is fully charged and discharged in each cycle. The author's experience of this meter is that functioning is more satisfactory with hard valves than with thyratrons because, with the latter, relaxation oscillations tend to occur. With the components given by Fig. 35 the author obtained full scale deflection for 1000 c/s, the minimum input voltage per transformer secondary to maintain linearity being 2 volts. Beyond this value the indication was independent of the magnitude of the input voltage. A stabilized H.T. supply must be employed and, if desired, a double triode, such as an ECC 33, may replace the two separate values.

A frequency meter, of the author's design, is shown by Figs. 36 and 37. This instrument employs hard valves and has the following outstanding characteristics :—

(*a*) Frequency range 0—50 Kc/s in six ranges.
(*b*) Independency of input voltage between 1·5 and 250 volts.
(*c*) Unaffected by non-sinusoidal waveforms.
(*d*) Input impedance 80K.
(*e*) Provision for checking calibration against mains supply.

Referring to Fig. 37, the primary function of the first valve is amplification of low input voltages in order to provide sufficient voltage on the second valve for squaring purposes. However, for inputs of more than a few volts saturation and cut-off limiting occur and the first valve functions as an overdriven amplifier.

Fig. 36.



C1 1μF	C7 0.005μF	R1 100K	R7 10K	R13 0.8K	V1 EF37	
C2 0.1μF	C8 0.001μF	R2 100K	R8 2.5MΩ	R14 —	V2 EL33	
C3 16μF	C9 0.0005μF	R3 220K	R9 5K	R15 —	V3 EB34	
C4 0.1μF	C10 0.0001μF	R4 3.3K	R10 1.5K	R16 3K	V4 5130	
C5 0.05μF	C11 16μF	R5 220K	R11 0.2K	R17 3.3K	V5 442BU	
C6 0.015μF	C12 16μF	R6 330K	R12 —	L 1.2H	M 0-1mA	

Fig. 37.

This latter property is not, however, essential and occurs only when the input voltage is in excess of 5 volts. With 12 volts, r.m.s., and above, applied to V_2, saturation and cut-off limiting occur in this valve and it generates an approximately square output voltage wave the amplitude of which is 110 volts. Thus when cut-off occurs the anode voltage becomes equal to that of the H.T. supply, 120 volts, while with saturation the anode voltage falls to 10 volts. These effects are shown by Fig. 38. It will be noted from Fig. 38 that the edges of the wave are not quite perpendicular. This effect is caused by the charging and discharging of C_7. When cut-off occurs the H.T. voltage is applied to C_7 *via* R_9 and the L.H. diode. Thus the leading edge of the wave is given by

$$e = E\,(1-\epsilon^{-\frac{t}{CR}})$$

$$\text{and } e/E = (1-\epsilon^{-\frac{t}{CR}}) \qquad (24)$$

During saturation, C_7 discharges through V_2, the R.H. diode and M. The resistance of the discharge circuit being lower than that of the charging circuit, C_7 is discharged more rapidly than it is charged. Hence considerations regarding charge and discharge times may be restricted to charging only.

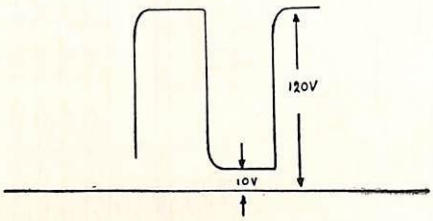


Fig. 38.

In order that M may have a linear scale it is necessary that C should be fully charged and discharged independently of frequency. Since this is theoretically impossible a criterion is taken that e/E in (24) shall not be less than 0·99, *i.e.*,

$$\epsilon^{-\frac{t}{CR}} \not> 0{\cdot}01$$

$$\text{or } \frac{t}{CR} > \log_\epsilon 100$$

$$i.e.\ \frac{t}{CR} > 4{\cdot}6$$

The charging time is approximately equal to $1/2f$ seconds, where f is the frequency undergoing measurement. Hence

$$\frac{1}{2fCR} > 4{\cdot}6$$

$$\text{or} \quad \frac{E}{2fCER} > 4{\cdot}6$$

where E is the value of the H.T. supply voltage.

But the charging and discharge currents are each equal to fCE amperes and thus

$$\frac{E}{2\,IR} > 4{\cdot}6$$

$$\text{or} \quad I < \frac{E}{9{\cdot}2R}$$

With E $\fallingdotseq$ 120 and R_9 = 5,000 this criterion is easily satisfied if an 0—1mA meter is employed. Since the meter current is equal to fCE amperes it is apparent that on any range CE must be rigidly constant. To meet this requirement a stabilized supply and mica capacitors are employed. The function of R_{11} is to neutralize the electron initial velocities in V_3.

Range and Calibration.

Six ranges are employed, namely, 0—150, 500, 1,500, 5,000, 7,500 and 50,000 c/s.

Since $I = fCE$

$C = I/fE$

or, with I equal to 1 mA,

$C = 10^3/fE$ microfarads.

and this expression gives the nominal value of the capacitance when the required full scale deflection is known. Thus, for f = 150, C = 0·055μF, while for f = 50,000, C = 0·000166μF. The values given by Fig. 37 were the nearest nominal ones available but capacitors were selected having values about 10% in excess of these. S_1 and S_2 are ganged, the function of S_2 being to provide a shunt for the meter when the capacitor value is in excess of that required. The meter has two scales and was precalibrated on the assumption that the relation I = fCE would strictly hold. Hence, with this assumption, it was necessary to check the ranges at a few points only. Thus the 150 range was checked against the mains frequency after first ascertaining that the latter was precisely 50 c/s. The 500 range was checked with a 400 c/s alternator the latter being positively driven by a synchronous motor operating from a precise 50 c/s supply. The 1500 range was checked at 400 c/s as previously and at 1,000 c/s by means of an oscillator and Wien's bridge. The remaining ranges were all checked, and adjusted where necessary, by means of the oscillator and bridge.

The function of S is to allow the meter to be switched to the mains supply for checking purposes. The signal voltage is derived from a 25 volt secondary winding, as shown. The instrument weight is 15 lbs.

Frequency-Meter Tachometer.

It is apparent that a frequency meter of the foregoing types could be employed as a tachometer. For example, a small a.c. generator could be fitted to the shaft of which the speed was desired and the output of this fed to the frequency meter. An alternative would be a generator, sufficiently small to be held, that could be engaged with the shaft by means of a rubber tip. Such generators could be of miniature dimensions because only a few volts are required with an almost negligible current. However, a tachometer of this type is not suitable for speeds below about 1000 r.p.m. because at this speed, flicker of the meter indicator may occur.

A further method of employing a frequency meter as a tachometer is to mount a small circular mirror on the shaft, one half of this mirror being blackened. A light beam is then projected onto the mirror and the reflected beam concentrated on a photocell. The output of the cell is of rectangular waveform and of frequency corresponding to that of the shaft rotation. Hence this output may be employed for operating the tachometer.

Stroboscopes, Tachometers and Flashlamps.

These instruments have been treated in detail by the author elsewhere* and, hence, here, will be discussed briefly only. Stroboscopes and tachometers generally comprise a power unit, a gas discharge lamp, and an oscillator for periodically triggering the lamp. In the smaller sizes the lamp is neon-filled, partly because of the distinctive colour of this gas. The discharge is of the arc type and the continuous loading of the lamp is about 5 watts. With a flashing frequency of 3,000 per minute this gives 0·1 joule per flash. The light output at this power is relatively small and, with a reflector of small divergence, results in an average illumination intensity of about 3 ft. cds. at $1\frac{1}{2}$ feet from the lamp. The discharge time per flash is approximately 100 microseconds.

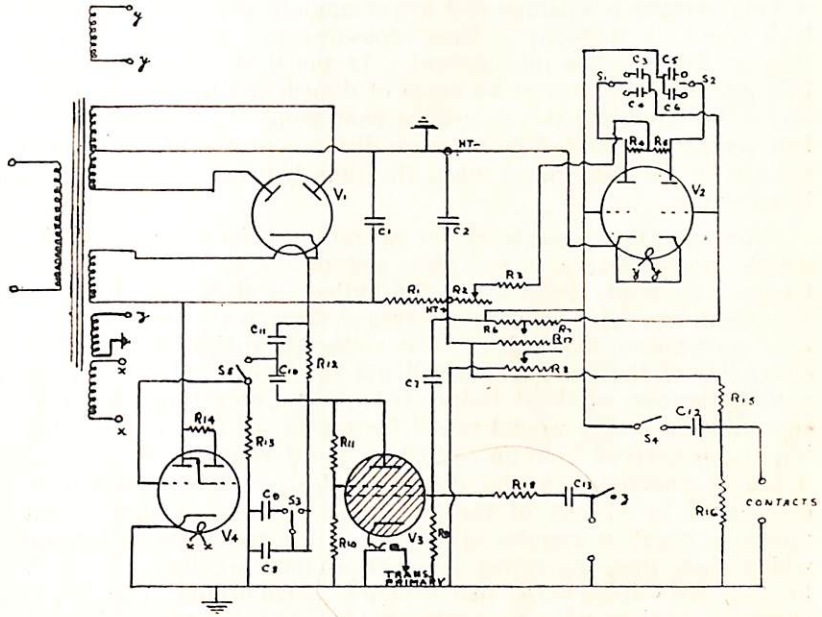
Triggering the lamp is effected by means of applying steep-fronted negative pulses to the lamp grid by means of a multivibrator or blocking oscillator. The frequency of the oscillator may be continuously varied over a range of about 5 : 1 and hence the flashing frequency may be varied in the same ratio. The

* *Electronic Stroboscopes, Tachometers and Flashlamps*, Association of Engineering and Shipbuilding Draughtsmen.

Considerations in the Design of Photoflash Equipment, Beama Journal, September, 1950, p. 264.

frequency is determined either by precalibration of the frequency—changing potentiometer or by means of one of the frequency meters previously described.

When illumination intensities greater than those given by a small neon lamp are required, a lamp of the type commonly employed for high-speed photography is used. These lamps have a gas filling of argon, krypton, zenon, or a krypton-zenon mixture, and the spectrum of the discharge approximates to northern daylight with a colour temperature of about 6500°K. Such lamps are, of course, very suitable for photographic work. The lamps



R1 3·5K 3W	R10 10K ±W	C1 8μF 500V	C10 0·0005μF 350V
R2 30K 1W	R11 50K ±W	C2 16μF 500V	C11 0·005μF 350V
R3 30K 1W	R12 2·5K 5W	C3 0·02μF 350V	C12 0·0005μF 350V
R4 100K 2W	R13 1M ±W	C4 0·005μF 350V	C13 0·1μF 350V
R5 100K 2W	R14 1M ±W	C5 0·02μF 350V	V1 43 IU
R6 1M 1W	R15 50K ±W	C6 0·0005μF 350V	V2 ECC 32
R7 1M 1W	R16 100K ±W	C7 0·0005μF 350V	V3 N6P 1
R8 1·5K 1W	R17 10K 20W	C8 1μF 350V	V4 Y 63
R9 1M ±W	R18 100K ±W	C9 2μF 350V	

Fig. 39.

require a much higher triggering voltage than is available from an oscillator and hence are triggered by means of the secondary winding of an oscillation transformer. The primary winding of this transformer is connected in series with the lamp of an ordinary neon-lamp stroboscope in the manner indicated by Fig. 39. Thus the larger lamp is fired each time the smaller lamp is triggered by the oscillator. The circuit of Fig. 39 is that of a stroboscope designed by the author, and for full details of this the reader is referred to the reference previously quoted.

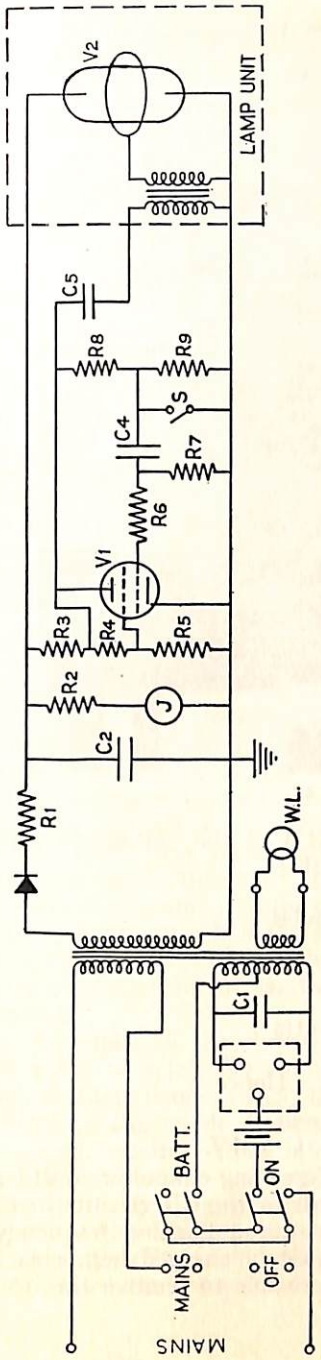
The lamps employed in high-illumination stroboscopes are usually termed flashlamps and are commonly employed for very high speed photography. Thus exposures may be made with a time as short as one microsecond. In the field of measurement this permits estimates to be made of dimensions of deformations under impact, while two successive photographs of a projectile at two positions separated by a known distance permit the projectile velocity to be determined when the time between the exposures is known.

The equipment necessary for operating a flashlamp is fairly simple, comprising a power pack and means for triggering the lamp. A circuit, designed by the author, is shown by Fig. 40. A capacitor of 50μF capacity is charged through the resistance R_1 to a potential of 2000 volts. This voltage is applied to the main electrodes of the lamp which will not fire, however, until a high voltage impulse, of about 15-kV., is impressed on a third electrode consisting of a wire wound round the outside of the lamp. This impulse is derived from an oscillation transformer the primary of which is energised by the discharge of C_5. This capacitor is discharged by closure of the switch S. Closure of this switch applies a negative impulse to the grid of the cold-cathode tetrode which then fires, providing a low resistance arc-discharge path for C_5. It will be noted that a meter, J, calibrated in joules, is employed to show when C_2 is fully charged, and that the equipment may be operated either from the mains or a self-contained accumulator. A photograph of the ebullition of water, taken by means of the equipment, is shown by Fig. 41. Fuller details may be obtained from the appropriate reference previously quoted.

The circuit shown by Fig. 40 is suitable for single-flash photographs only. However, it is apparent that if S is periodically operated a series of flashes will result and a succession of photographs may be taken.* For this type of work, however, a far smaller capacitor than 50μF must be employed, otherwise the lamp will be overloaded and C_2 will not be fully charged in the intervals

* *An Inexpensive Stroboscope for High Speed Photography*, Shirleigh Silverman and W. H. Warhus, Rev. Sci.Insts, September, 1943, p. 273.

A High-power Stroboscope, D. A. Senior, J. Sci.Insts., April, 1946, p. 81.



R1	20K	50W
R2	1MΩ	
R3	2·5MΩ	2½W
R4	1MΩ	½W
R5	0·22MΩ	½W
R6	0·1MΩ	½W
R7	1MΩ	½W
R8	1MΩ	½W
R9	0·22MΩ	½W

C1	1μF	350 V.W.
C2	50μF	2000 V.W.
C4	0·1μF	350 V.W.
C5	1·0μF	350 V.W.

V.1	ASP 1
V2	SF 3

W.L.	2·5 VOLTS 0·2 AMPS
VC	WEARITE NS4

Fig. 40.

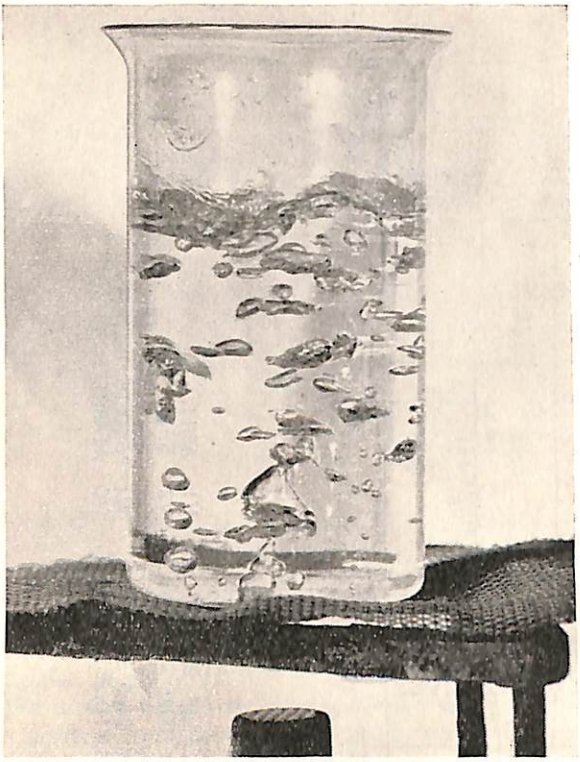


Fig. 41.

between successive discharges. If V is the voltage to which C_2 is charged then the energy per flash is

$$E = \frac{1}{2} CV^2 \quad \text{joules}$$

and the average power is

$$W = \frac{1}{2} fCV^2 \quad \text{watts}$$

where f is the flashing frequency. Hence

$$C = 2W/fV^2$$

and, for given values of W and V, $C \propto 1/f$.

For photographic work of the foregoing character consideration must be taken of the time-constants of the CR circuits in Fig. 41 in order to ensure that, at the required flashing frequency, the capacitors C_2, C_4 and C_5 are adequately charged before each discharge. Also it may be found desirable to remove the iron core

from the oscillation transformer in order to increase the rate of discharge of C_5. In practice S consists of an accurately constructed contactor, mounted on the film driving sprocket. The camera film runs at a high speed past the lens, the latter remaining open while the film is in motion. The speed of the film is invariably so high that normal indoor light, received through the lens, is insufficient to have any effect. It is apparent that a series of photographs, taken in the above manner, permit the measurement of projectile velocities. Thus if two images are obtained at a known distance apart then, with a knowledge of the interval between exposures, the velocity of the object may be deduced.

Time Measurement and Control.

For the purposes of time measurement and control, electronic methods are mainly applicable to relatively short intervals, generally not exceeding 100 seconds. For longer intervals mechanical methods and synchronous motors become advantageously competitive owing to their greater accuracy. Two types of electronic timers may be distinguished. One type measures a time interval, in a similar manner to a chronometer, while the other indicates the lapse of a predetermined time interval or controls the time of some operation. The first type is usually referred to as a timer or chronometer and the second type as a process timer. Typical applications of the first type are measurements of pull-in and drop-out times of circuit breakers, times of flight of projectiles, camera shutter times, etc., while applications of the second type are found in welding, high-frequency heating, photographic enlarging, etc. Electronic timers usually employ a capacitance-resistance circuit in such a manner that the time interval concerned is equal to that required for the capacitor voltage to change from one value to another. With timers conditions are arranged that the change in values is proportional to the time for the change. Hence, appropriate measurement of the change in values measures the time. With process timers, when the second value is reached indication of this fact is given by some form of indicator, the action of the indicator being dependent on the properties of either a hard valve or thyratron.

The principles of the electronic process timer may be demonstrated with the relatively simple circuit of Fig. 42. This timer employs a diode-triode valve, the diode section for charging the capacitor and the triode section for holding off the relay current until a predetermined time interval has elapsed. The cycle of operations is as follows : On pressing the push button, P.B., the capacitor is charged through the diode section, the upper plate of the capacitor thus acquiring a negative potential. This potential then biasses the triode, cuts off the relay current and hence either opens or closes the relay contacts, according to the arrangement

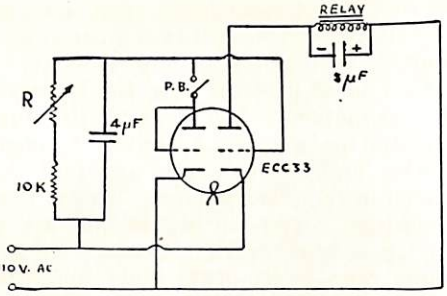


Fig. 42.

of these. When the capacitor is fully charged, the push button is released, whereupon the capacitor commences to discharge through R. This reduces the bias on the triode section and, when the voltage has fallen to the cut-off value, current recommences to flow in the relay thus actuating the contacts. The time interval is, of course, that occurring between release of the push-button and operation of the relay, this interval being predetermined by the setting of R.

A somewhat similar circuit to that of Fig. 42 is shown by Fig. 43. However in the latter circuit a thyratron is employed in place of hard valves and this serves the double purpose of providing half-wave rectification for charging the capacitor and, later, operating the load or relay. With S open the capacitor is charged through the thyratron, the grid acting as an anode. Thus the upper plate of C acquires a negative charge. On closing S, the full potential difference of the transformer primary appears between anode and cathode but the p.d. between grid and cathode is now that across the capacitor. Hence the thyratron cannot fire until the capacitor

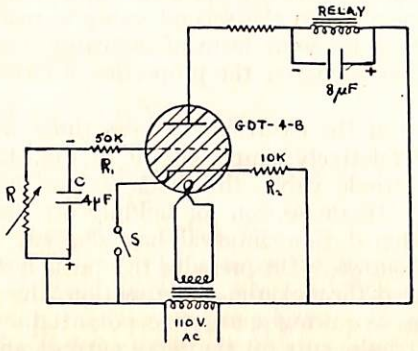


Fig. 43.

voltage falls to the critical grid voltage. Repetition of the cycle is obtained by opening S.

The circuits illustrated by Figs. 42 and 43 have several defects that make them unsuitable for precise timing. For example, during charging, the presence of R across C does not allow the capacitor to be fully charged owing to the anode-cathode drop in the diode and the drops across the resistances R_1R_2 in Fig. 43. This effect is most marked for low values of R and is particularly apparent with the circuit of Fig. 42. Should the closure of P and S be too brief, again the capacitor is not fully charged. Variations in the supply voltage cause the value of the triode cut-off voltage to change and thus modify the timing interval. Such variations also modify the voltage to which C becomes charged.

The Discharge Characteristic.

During discharge the circuit equation of the RC combination is

$$Ri + e = 0$$

where i and e are respectively the capacitor voltage and current. Since $i = Cde/dt$

$$CR\frac{de}{dt} + e = 0$$

$$\text{or} \quad \int \frac{de}{e} = -\frac{1}{CR}\int dt$$

$$\text{and} \quad e = E_o\, \epsilon^{-\frac{t}{CR}}$$

where E_0 is the capacitor voltage at the moment at which the discharge commences. If E_c is the triode grid voltage at which the relay operates, or alternatively, the critical grid voltage of the thyratron, then

$$t = CR \log \frac{E_o}{E_c} \tag{25}$$

and this expression may be written

$$t = CR \log K$$

where, for a thyratron, K is the control ratio. For a hard valve, K is the ratio of the anode voltage to that value of the grid voltage at which just sufficient current flows to operate the relay. For both valves K is not constant.

In these equations t is the time interval between the commencement of the discharge and the capacitor voltage reaching the value E_c. Since t is proportional to R, the timing interval may be varied by variation of R. It is apparent that any variation in E_o and E_c will affect the timing interval unless these quantities

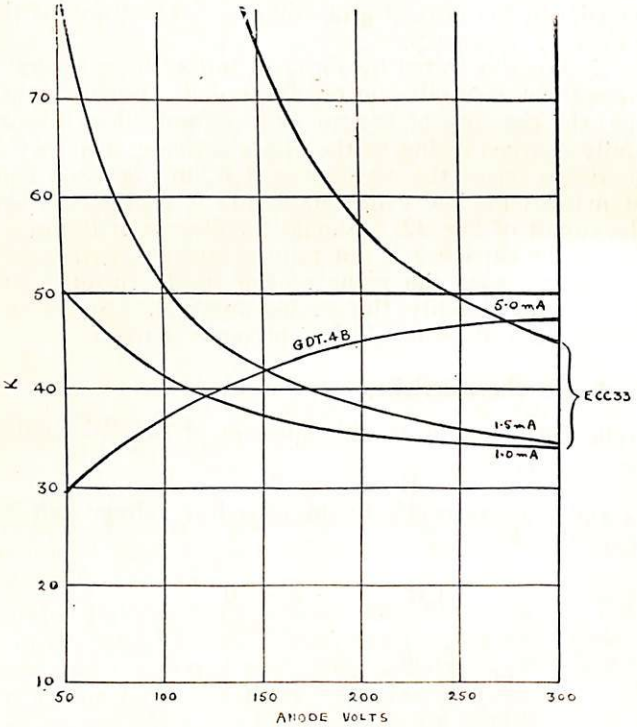


Fig. 44.

simultaneously vary in the same ratio. As previously stated, K is not constant and hence variations in E_o and E_c affect the value of the timing interval. Fig. 44 shows the variation in K for GDT.4B and ECC 33 valves. It will be noted in both cases that the rate of variation of K is smaller at high anode voltages. Also, for the ECC 33, the rate of variation is smaller for low anode currents. Hence stability of the timing interval is favoured by high anode voltages and, in the case of hard valves, by low relay actuating currents. Of course the change in timing interval is not as great as the curves might suggest because the rate of change of log K is considerably less than that of K. Consideration of the curves suggests that an increase in anode voltage increases the time interval with a thyratron circuit and decreases the interval with a hard valve circuit. Experiment supports this suggestion.

For precision interval timing the author has designed the process timer of Figs. 45 and 46. It will be noted that a stabilized supply is employed thus rendering the intervals independent of mains voltage variations. The other outstanding feature is that

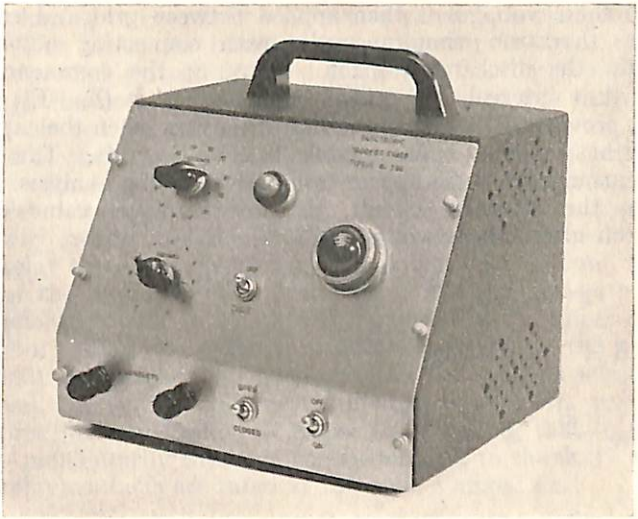
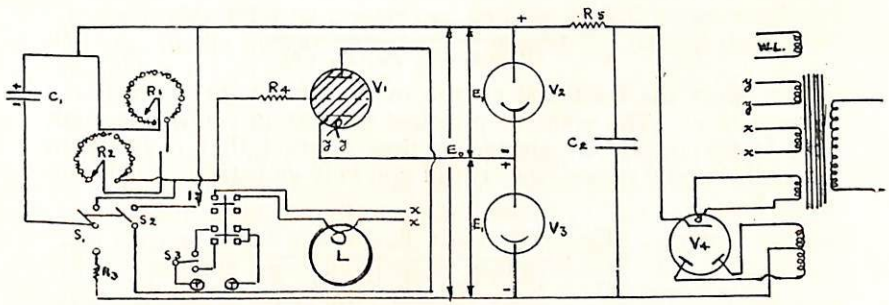


Fig. 45.



R1 11-5 M 1W	C1 3·5 μF
R2 10-0·5 M 1W	C2 8 μF
R3 22K ½W	V1 GDT-4-B
R4 100 K 1W	V2 S130
R5 3K 5W	V3 S130
L 12V 10W	V4 EZ 35

Fig. 46.

having charged the capacitor to a voltage $E_o \fallingdotseq 2E_1$, the capacitor voltage is then connected in opposition to a voltage E_1 and the sum of these voltages is then applied between grid and cathode of the thyratron simultaneously with connecting across the capacitor the discharge resistor. Thus, at the commencement of the time interval, the grid-cathode voltage is $(E_o - E_1) \fallingdotseq E_1$. Hence, provided E_c is small, the thyratron fires when the capacitor voltage has fallen to approximately $E_o/2 \fallingdotseq E_1$ volts. The merits of this arrangement will appear from the following analysis. Considering the discharge circuit, the instantaneous value of the thyratron grid-cathode voltage is

$$e = E_o \epsilon^{-\frac{t}{CR}} - E_1$$

$$\text{and } t = CR \log_\epsilon \frac{E_o}{E_1 + e}$$

$$\text{Also } \frac{dt}{de} = -\frac{CR}{E_1 + e}$$

$$\text{Now } e/t = \frac{e}{CR \log E_o/(E_1 + e)}$$

$$\text{and } \frac{dt}{de} \cdot \frac{e}{t} = \frac{dt/t}{de/e} = -\frac{CR}{E_1 + e} \times \frac{e}{CR \log E_o/(E_1 + e)}$$

$$\text{or } \frac{dt}{t} = \frac{de}{e} \cdot \frac{e}{(E_1 + e) \log E_o/(E_1 + e)} \qquad (26)$$

which gives the fractional change in time for a given fractional change in e. The point of principal interest in connection with (26) is the percentage change in time interval that is likely to occur if, for any reason, the critical grid voltage varies. Replacing e by E_c,

$$\frac{dt}{t} = \frac{dE_c}{E_c} \cdot \frac{E_c}{(E_1 + E_c) \log E_o/(E_1 + E_c)} \qquad (27)$$

and the corresponding equation for circuits of the types shown by Figs. 42 and 43 is obtained by putting $E_1 = 0$. Now if $E_c \ll E_1$ and $E_o = 2E_1$ (27) becomes

$$\frac{dt}{t} = \frac{dE_c}{E_c} \cdot \frac{1{\cdot}44\ E_c}{E_1}$$

For the circuit of Fig. 46, $E_c \fallingdotseq 3$ volts and $E_1 = 120$ volts. Thus a 33% variation in E_c results in a 1·2% variation in the time

interval. With the circuits of Figs. 42 and 43 the variation in timing would be 7·5%.

Referring to Fig. 46, it will be noted that C_1 is charged through R_3. This limits the initial charging current to about 10*m*A, at the same time permitting rapid charging. Discharge of C_1 occurs through the decade resistors R_1R_2 which permit any time interval to be selected, in one second steps, from 0 to 120 seconds. S_1 and S_2 are mechanically coupled but S_1 makes slightly in advance of S_2 in order to establish the bias on V_1 before the anode circuit is completed through the relay coil. S_1S_2 is an ordinary double pole two-way switch, the side closing first being determined by trial. The relay serves two purposes. It switches on the lamp L at the end of the required interval and either opens or closes a pair of contacts according to whether S_3 is placed in the upper or lower position. Thus an external circuit, connected to the terminals TT, may be switched either on or off at the end of a selected time interval. When a cycle is terminated recycling is very rapid, the time interval that must elapse being merely that to recharge C_1 by momentarily throwing the switch S_1S_2 to the "off" position. The relay contacts are rated at 250 volts 5 amps. a.c.

Discharge Circuit.

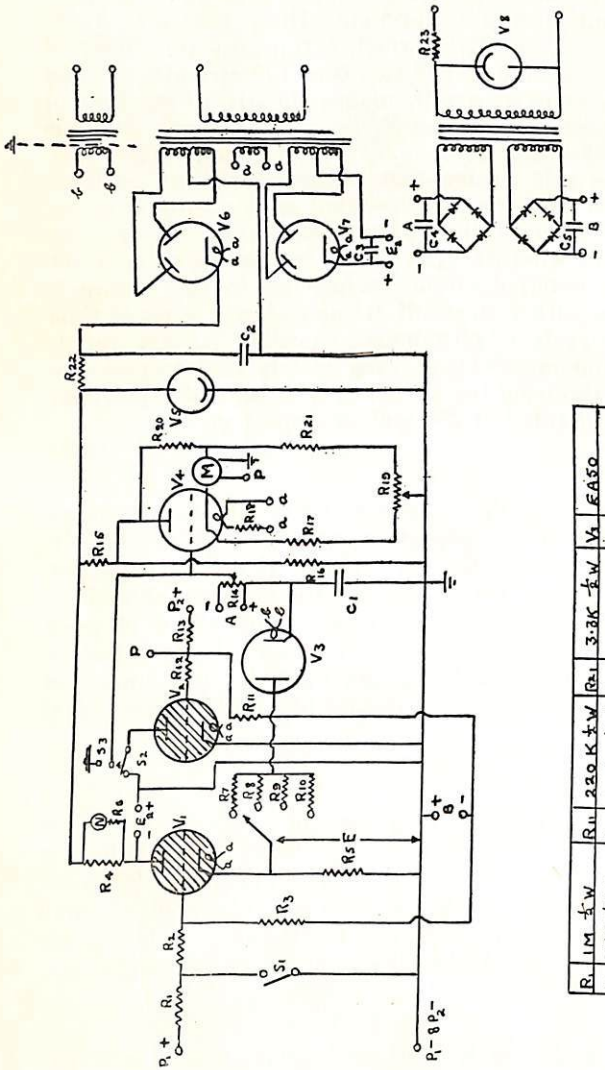
For precision interval timing the stability of the discharge circuit is, of course, equally important with that of the supply voltage and the firing point of the thyratron. Hence the capacitor must have high insulation resistance and the resistors should be of the high-stability cracked-carbon type. The effect of poor capacitor insulation resistance becomes progressively effective as the timing interval is increased and for the RC values employed in Fig. 46 the insulation resistance should not be less than 1,000 megohms.

Calibration.

Calibration may be effected by means of a good stop watch and adjustment of C_1. Thus, if the precise value of C_1 cannot be obtained in a single unit, a slightly smaller value should be taken and then adjusted upwards by the addition of a minimum of small units until the desired timing accuracy is attained. The instrument of Fig. 45 is accurate within 1·0% over the entire range.

Timers.

Timers are invariably concerned with smaller time intervals than are the previously described instruments. A timer, of the author's design, for measuring time intervals from 0·2 millisecond to 1 second, is shown by Figs. 47 and 48. The basic principle of this instrument is that a capacitor is charged at constant current



R	1M ±W	R11	220K ±W	R21	3·3K ±W	V1	EA50
R2	15K ±W	R12	10K ±W	R22	3K 5W	V4	EF37
R3	1M ±W	R13	220K ±W	R23	5K 5W	V5	4687
R4	10K 5W	R14	22K 3W	C1	1μF	V6	EZ35
R5	3·3K 1W	R15	15K 1W	C2	8μF	V7	EZ35
R6	47K ±W	R16	10K 1W	C3	8μF	V8	4687
R7	200K ±W	R17	3·3K ±W	C4	50μF	M	0-100μA
R8	1M ±W	R18	15Ω ±W	C5	50μF		
R9	4M ±W	R19	1K 3W	V1	2D21		
R10	18M ±W	R20	147K ±W	V2	2D21		

Fig. 47.

Fig. 48.

during the time interval to be measured. If t is this time and I is the current, then,

$$It = Q = eC$$
$$\text{and } t = eC/I$$

where C is the capacitance and e is the voltage rise on the capacitor terminals during t. Since t is proportional to e, measurement of e determines t.

It is apparent that means must be provided for starting and stopping I, without delay, at the commencement and termination of the time interval. This is effected by means of a pair of thyratrons which constitute an electronic switch. Application of a positive impulse to the grid of V_1 fires this valve and causes a voltage to appear across R_5 in the cathode lead. This voltage persists until a second positive pulse, at the end of the time interval, fires V_2. When V_2 fires, the anode-cathode voltage of V_1 abruptly changes from E_a to $-(E_2-E_a)$ thus causing extinction of the latter valve. Firing of V_1 and V_2 is particularly rapid, being less than 1 microsecond after application of a positive pulse. The time interval measured is that elapsing between the two pulses and hence external arrangements must be such that these pulses may be obtained from the phenomenon undergoing measurement.

The capacitor charging current is derived from the voltage E appearing across R_5. Thus C_1 is charged through R_7 and the diode,

V_3, by means of E, and the instantaneous value of the charging current is

$$i = C_1 \frac{de}{dt}$$

where $e = E\,(1 - \epsilon^{-\frac{t}{CR}})$

e being the instantaneous value of the capacitor voltage. The current may also be written as

$$i = \frac{E - e}{R_5} = I - \frac{e}{R_5}$$

and, provided $E \gg e$, $i \fallingdotseq I$. In the present instrument E is 15 volts and the maximum value ever attained by e is 0·7 volt, the latter figure corresponding to full scale deflection of the valve voltmeter. Hence over the time interval to be measured de/dt is approximately constant and so also is I.

Thus $e \propto t$

$$\text{and } t = \frac{e\,CR}{E} \qquad (28)$$

It is apparent that the range of the instrument may be varied by variation of either C or R or both. The voltage drop across the diode during charging is negligibly small, the function of this valve being to prevent the capacitor charge from leaking. Evidently leakage caused by anything connected to the capacitor is of the greatest importance and hence extremely high values of insulation resistance are necessary. That of the capacitor is 2000 megohms. The capacitor will, of course, tend to charge owing to the effects of electron initial velocities in the diode and to suppress these effects an adjustable backing e.m.f. is employed. This e.m.f. must be stabilized, otherwise voltmeter drift results. Details of the method of stabilization are given on p. 101.

Operating the diode heater and the backing e.m.f. from normally-wound transformers, it was initially found that considerable leakage occurred and that a relatively high backing e.m.f. was required. These difficulties were overcome by employing separate, screened, transformers for these operations, the transformers being constructed to have a large interwinding resistance and a high secondary winding to core resistance. Slight leakage of the charge from C_1 is of course bound to occur, but it must be sufficiently small so that no difficulty is experienced in obtaining accurate readings.

Voltmeter.

The voltage acquired by C_1 is measured by means of a degenerate voltmeter of the type described on p. 29. To avoid leakage, V_4

is operated with an open grid, this being made possible by operating the heater at 4 volts and the anode at 28 volts. Full scale deflection occurs for 0·7 volt. The scale is marked in milliseconds and from (28) and the theory of these voltmeters it is apparent that the angular deflection of the indicator is proportional to time. In the event of the time interval between the pulses being greater than that corresponding to full-scale deflection the meter will tend to be overloaded. However, the smallest capacitor charging resistor has a value of 200K and thus limits the maximum possible grid current of V_4 to 15/200K = 75 μA. The flow of grid current, of course, limits the voltage that can be applied to the grid of V_4 and this and the degeneracy of the valve-voltmeter circuit adequately protect the meter from damage.

Once the second pulse has occurred, it is necessary momentarily to break the anode circuit of V_2 before a further measurement may be made. This is effected with the push button S_2 which, when pushed far enough, also removes the charge acquired by C_1 from E. The neon lamp shunted across R_4 indicates whether V_2 is conducting or not. The bias switch S_1 permits testing the timer by tripping V_1.

Fitted to the meter is a pair of contacts. One of these is stationary while the other is attached to the meter movement. Just beyond F.S.D. the contacts meet and, by so doing, fire V_2. Thus when V_1 is tripped by means of S_1, V_2 is automatically tripped by the meter. The latter action also occurs when a second positive pulse fails to arrive at P_2+.

Calibration.

In order to calibrate the timer it is necessary to have some means of producing pairs of pulses with an accurately known separation. For the present purpose transient pairs were produced by means of the circuit of Fig. 49. The first pulse, P_1, is produced the moment S is closed and the second pulse, P_2, the moment the

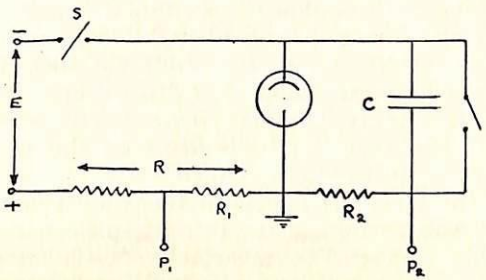


Fig. 49.

capacitor voltage reaches the striking potential of the lamp. The voltage on C is given by

$$e = E\,(1-\epsilon^{-\frac{t}{CR}})$$

from which $$t = CR \log \frac{E}{E-E_s}$$

where E_s is the lamp striking voltage. Hence transient pairs are produced with a known separation, t, provided C, R, E and E_s are accurately known. The magnitude of these pulses is given by $P_1 = R_1E/R$ and $P_2 = IR_2$ where I is the initial value of the capacitor discharge current. Variation of t can, of course, be obtained by varying R.

It will be noted that four ranges are employed corresponding to maximum values of 10, 50, 200 and 1,000 milliseconds.

Insulation Testing.

The failure of insulating media under electric stress may be due to various causes. Among these are chemical and physical conditions and the way in which the stress is applied. Since there is no perfect insulator, electric stress in an insulating medium invariably results in a flow of current, small but definite, and this current is generally electronic in nature. However, owing to certain types of unwanted inclusions, including moisture,* current flow may be electrolytic in nature and, because of detrimental effects of electrolysis, failure of the insulator may occur. If the inclusions in an insulator take the form of gas-filled cavities discharges may occur in these. When these discharges are sustained they may lead to ultimate failure of the medium. The discharges tend to enlarge the cavities and the discharge products cause chemical changes and carbonization. That the manner in which the stress is applied may lead to premature failure is fairly well-known. Thus stress concentration at sharp corners and edges will lower the breakdown strength or cause a discharge in the adjacent insulating medium.

In some instances breakdown of an insulator may occur through what is termed, "thermal instability." This means that the dielectric losses resulting from the stress are such that the rate of production of heat exceeds that of dissipation. It is apparent that thermal instability is much more likely when the stress alternates and also that it is dependent on the shape, size and ambient temperature of the specimen.

When all the foregoing causes of breakdown are absent there remains what may be termed the intrinsic dielectric strength of a medium. This strength is invariably far greater than that

* *Aircraft Electrical Engineering*, F. G. Spreadbury (Pitman), 2nd Ed., p. 123.

experienced in practice because some, if not all, of the above-mentioned phenomena are always present in an insulating medium. Intrinsic breakdown is purely electronic in nature and further details of this process may be found in the reference below.*

From foregoing statements it is apparent that when practical insulation tests are made it is seldom the intrinsic strength that is tested. Rather the tests determine to what extent undesired factors lower this strength or are likely to shorten the life of the insulation. When cavities are present the discharge inception voltage is, generally, far less than the practical breakdown voltage and hence, provided suitable test equipment is available, tests for cavities may be made at voltages which are low relatively to that of breakdown. The presence of cavities is particularly important in impregnated apparatus because they indicate the extent to which impregnation has been successful.

The desirability of non-destructive methods of testing is scarcely to be questioned, particularly when the test subject is of complex or costly structure. In the electrical field the process of flash-testing insulation materials has frequently been criticized and here it is pertinent to quote some remarks due to Evershed.'† "To conduct tests without any means for ascertaining what is going on in the insulator as the breakdown voltage is approached, without observing the current or, better still, the resistance, is to shut our eyes and deliberately avoid looking for the cause of failure." However, the possibility exists of determining the condition of an insulator without loading it to the point of virtual destruction. A method of effecting this is to produce a gradually increasing d.c. potential gradient within the insulator while providing means for detecting the inception of ionization and determining its magnitude. In an insulator tested by this method thermal instability is negligibly small and, initially, the method reveals discharge inception and the presence of moisture. If the gradient can be made sufficiently high, excessive stress concentrations can be revealed and, ultimately, the practical breakdown strength can be measured.

The method of detecting discharges within an insulator, stressed by a steady d.c. potential, is based on the fact that the discharges and discharge currents are irregular in time. Hence a resistance placed in series with the d.c. supply experiences continuously fluctuating currents. These currents produce a frequency spectrum ranging from about 100 to 10,000 c/s and hence may be detected by means of a narrow-band amplifier. Thus a method of detection is to connect the input of the amplifier across a suitable

* *Intrinsic Dielectric Breakdown in Solids*, H. Frolich and J. H. Simpson, p. 185, Advances in Electronics (Academic Press).

† *The Characteristics of Insulation Resistance*, J.I.E.E., Vol. 52, No. 224 (15th December, 1913), pp. 51-83.

series resistor, the output of the amplifier being indicated by either a meter or monitor speaker.

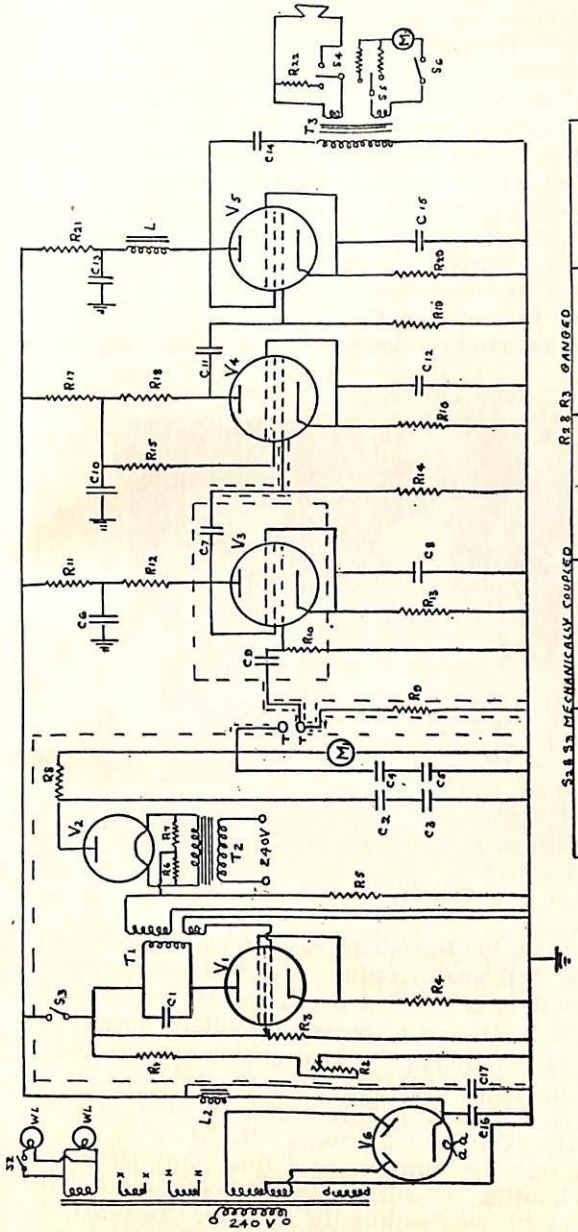
Nature of Discharges.

Considering a small spherical cavity within a dielectric (*e.g.*, caused by imperfect impregnation), an applied electric field, X, creates a uniform field within the cavity equal to $X(K+2)/3$,* where K is the dielectric constant. On opposite sides of the cavity connecting the field lines, surface charges of opposite signs appear, these charges being partly responsible for the field. When a discharge occurs electrons travel to the positive surface charge thereby partially neutralizing this. The neutralized area, being surrounded by an undischarged area, causes a transverse potential gradient under the influence of which transverse discharges may occur, thus further discharging the cavity surface. Similar effects, due to the flow of positive ions, occur at the negatively-charged surface.

From the foregoing argument it appears that as the voltage across a dielectric is raised a value will be reached at which discharge inception suddenly occurs. Maintaining the voltage somewhat above the inception value, discharges, should, initially, be continuous, afterwards progressively declining as the cavity surfaces become discharged. When discharges cease, raising the voltage should result in further discharges which, however, will ultimately decline as before. Should the voltage be sufficiently raised, the cavity charges may be reversed. Reduction of the voltage should then admit the possibility of reversed discharges. All the suggested phenomena actually occur and may be readily detected with the apparatus of Fig. 51. For example, when testing a $0{\cdot}1\mu F$ paper capacitor discharge inception commenced at 300 volts and, maintaining the voltage slightly above this value, the discharge ceased after about 20 seconds. Raising the voltage then caused a recommencement of the discharge which ultimately died away as before. After raising the voltage to 1600 and waiting for the discharge to cease, re-inception commenced when the voltage was lowered to 800. On lowering the voltage to zero and then reapplying, inception did not reoccur until the voltage reached 700. The last phenomenon is probably due to the field caused by the reversed cavity charges opposing that due to the applied voltage.

It will be appreciated from the foregoing discussion that an apparatus for demonstrating discharge effects in solid insulators must comprise a variable d.c. voltage and a suitable amplifier. The circuit of an insulation tester, designed by the author, is shown by Fig. 50, the complete equipment being shown by Fig. 51. Referring to the left-hand side of Fig. 50, it will be noted that

* *Electronics*, p. 138.



S1 & S2 MECHANICALLY COUPLED — R1 & R3 GANGED

C1 0·01μF 1000V	C11 0·005μF 350V	R1 6·2K 10W	R11 10K ½W	R21 3K 5W	
C2 0·01μF 1000V	C12 50μF 50V	R2 22K 3W	R12 22K ½W	R22 3Ω 1W	
C3 0·01μF 1000V	C13 8μF 500V	R3 ·22K 3W	R13 0·47K ½W	V1 EL 33	
C4 0·01μF 1000V	C14 0·1μF 350V	R4 0·4K 1W	R14 1M ½W	V2 EY 51	
C5 0·01μF 1000V	C15 50μF 50V	R5 5M 2W	R15 1M ½W	V3 EF37A	
C6 16μF 500V	C16 8μF 500V	R6 0·1K 5W	R16 33K ½W	V4 EF37	
C7 0·001μF 350V	C17 32μF 500V	R7 0·1K 5W	R17 47K ½W	V5 EL33	
C8 50μF 50V		R8 1M 1W	R18 220K ½W	V6 EZ35	
C9 0·001μF 350V	L1 1H	R9 47K ½W	R19 1M ½W	M1 0-2000V	
C10 0·001μF 350V	L2 6H	R10 1M 1W	R20 0·47K ½W	M2 0-2-10V	

Fig. 50.

the d.c. voltage is derived from a high-frequency oscillator, after rectification by a half-wave rectifier. The oscillator frequency is 35 Kc/s and the output voltage has stepless variation by means of variation of the screen voltage of the oscillator valve. By this means the d.c. voltage may be varied from 70 to 2,000 volts. As the screen voltage is reduced, so also is the valve current. Since the anode supply voltage is not stabilized, this causes the anode voltage to rise, and this, in turn, tends to limit the lower value to which the d.c. voltage can be brought. In order to give a degree of stabilization to the supply voltage, the resistance R_2 was introduced. This acts as a variable bleeder, the shafts of R_2 and R_3 being ganged. Thus, as the screen voltage and anode current are decreased, so also is the value of R_2. This is effected in such a manner that the total demand on the supply is roughly constant.

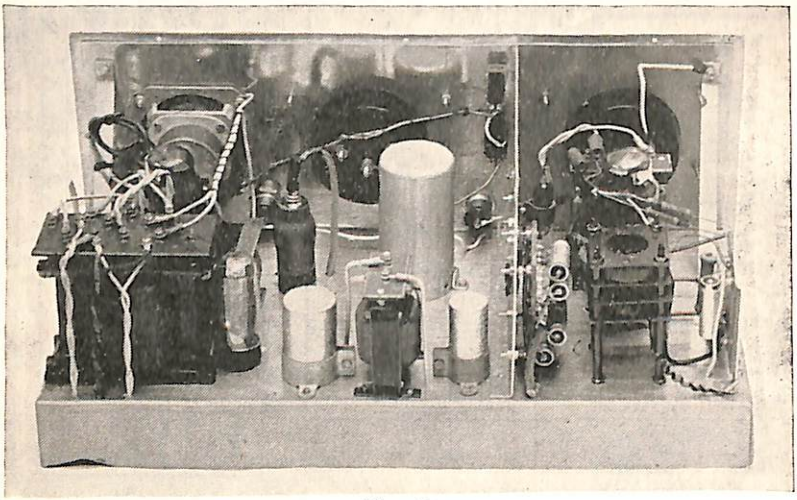


Fig. 51.

The oscillator unit is on the right-hand side of Fig. 51 in which the air-core transformer and small rectifier valve will be noted.

The output voltage of T_1 is rectified by V_2 and then smoothed by C_2 C_3, R_8 and C_4 C_5. It is, of course, absolutely necessary that no signal is fed to the amplifier but that arising from random currents in the insulator under test. It is for this reason that a h.f. oscillator is employed because a high degree of smoothing is easily effected by relatively small capacitors. R_8 has a value of 1 megohm and, apart from its function as a smoothing element, serves the purpose of limiting the current in the specimen undergoing test. It also prevents overloading the rectifier. In practice the regulation characteristic of the d.c. supply is such that the

current cannot exceed 0·2 *m*A. The effect of a complete breakdown is to cause the oscillator to stop operating and this, of course, completely safeguards the equipment.

Rectifier Heater Supply.

Although excellent filtering is obtainable by employing a h.f. supply, care must be taken to avoid the introduction of l.f. into R_9. With the diode connected in the manner shown, the upper test terminal is negative. If the diode is reversed this reverses the polarity of the upper terminal. This, however, brings the heater transformer into contact with the d.c. side of the system in the manner shown by Fig. 52. Here it will be noted that, owing to the interwinding capacitance of the heater transformer, a 50 c/s signal is impressed on R_9 via the capacitance of the subject undergoing test. The magnitude of the voltage appearing on R_9 depends on the values of the interwinding capacitance, the capacitance of the subject, and R_9. With an ordinary transformer

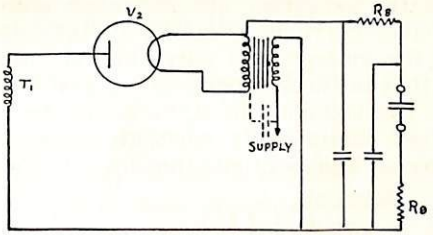


Fig. 52.

and a capacitance of 0·25μF between the test terminals the l.f. voltage appearing on R_9 is inconveniently large. This difficulty may be overcome by means of a heater transformer having an earth screen or by connecting the diode in the manner shown by Fig. 50. However, even with the latter method of connection the interwinding and earth capacitance should be low for otherwise a capacitive load is imposed on the oscillator. The heater transformer, T_2, is provided with screened windings in order to reduce inter-winding capacitance to a minimum.

Screening.

Although a h.f. supply may be readily smoothed, care must be taken to ensure that it does not affect the amplifier, either by capacitive or inductive coupling or by direct radiation. Such possibilities are avoided by completely screening the oscillator section in the manner indicated by Figs. 50 and 51. From Fig. 51 it will be noted that a screen is interposed between the oscillator and amplifier, the screening being completed by a top plate (not shown) and two sides of the case.

The Amplifier.

The voltage set up across R_9 by the random currents in the insulator undergoing test is amplified by a 3-stage, narrow-band amplifier having a gain of 3×10^4 or about 90 dbs. With a gain of this magnitude the necessity of keeping R_9 free from spurious effects is evident. The loudspeaker is, of necessity, mounted close to the amplifier and, initially, microphonic effects were experienced. These effects were eliminated by providing V_3 with a rubber mounting, using a highly flexible connector to the top cap grids of V_3 and V_4, and changing V_3 from an EF 37 type to an EF 37A. Despite the oscillator screening, a small amount of pick-up was experienced and to eliminate this it was found necessary to screen V_3. The values of the coupling condensers C_7, C_{11}, C_{14} are relatively small in order to attenuate the low frequency components, these components being due, mainly, to the power supply. The gain of the amplifier, expressed in both decibels and voltage ratios, is shown as a function of frequency by Fig. 53. Since the effects of dielectric ionization are indicated both visually and aurally, the decibel curve is primarily of interest in the latter connection and the voltage gain curve in the former connection. It will be noted that the decibel curve is reasonably flat from 200 to 12,000 c/s while the voltage-gain curve is also from about 300 to 10,000 c/s. These limits easily cover the noise spectrum while attenuating spurious signals originating from the power supply.

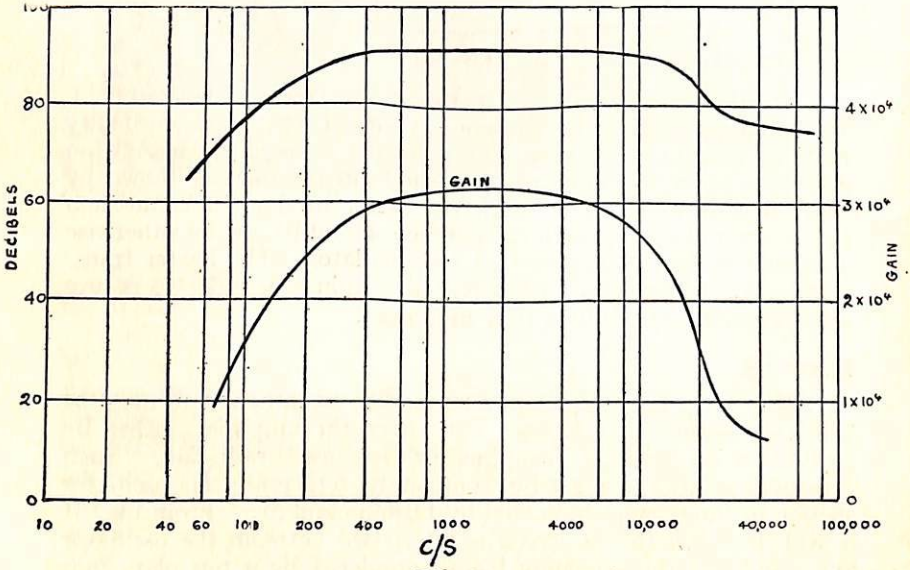


Fig. 53.

By connecting the first valve as a pentode it is possible to increase the amplifier gain by a factor of about 6. However, this increases the possibility of microphonic effects and also the susceptibility to pickup of extraneous signals, while only increasing the decibel gain by 18%. The susceptibility to external pickup is likely to be marked when components of large physical dimensions are tested, because such components are apt to act as aerials.

As shown by Fig. 50, indication of ionization is given by means of a meter and monitor loudspeaker. The speaker speech coil impedance is 3·0 ohms and handles 1·0 watt at full load. This loading occurs for 0·06μA in R_9. When aural indication is not required, a 3·0 ohm resistance is substituted for the speech coil. The meter is of the rectifier type and has two ranges marked 0·1μA and 0·02μA. Waveform errors exist with this type of meter but, since it is comparative rather than absolute measurements that are made, this type of meter is sufficiently satisfactory for this purpose. Owing to the random nature of the ionization current, sudden fluctuations of the meter needle tend to occur. These are minimized by providing the meter with particularly heavy damping.

The weight of the instrument is 31 lbs.

The Cathode Ray Oscillograph.

In the introduction it was stated that the c.r. oscillograph is one of the most important of all electronic instruments. Its importance is such that nothing less than a separate treatise can do justice to this subject, and, of course, many excellent works exist thereon. Here it is not possible to do more than draw attention to its importance and to recommend those interested to refer to some specialized work. The employment of c.r. tubes in television and radar is, of course, well-known and, in addition, it may be stated that it may be employed as a measuring instrument in connection with the following :—voltage, current, power, power factor, frequency, time, waveform analysis, iron losses, dielectric losses, etc., etc. As an introduction, the reader may consult the author's pamphlet on the subject* and also Chapters VI and XVII of his book, *Electronics*.

Applications of Electronics to Bridges.

Fundamentally, bridges are not electronic devices but their scope and usefulness have been greatly increased as a result of the application of electronics thereto. Electronics is usually applied to a bridge in two ways (*a*) as a power source ; (*b*) as some form of detector. Originally a.c. bridges employed a 50 c/s supply

* *An Introduction to the Cathode Ray Oscillograph*, Association of Engineering and Shipbuilding Draughtsmen.

as a power source or some form of oscillation transformer. The latter source enabled the null-point to be determined with greater accuracy but, for some purposes, had the disadvantage of a non-sinusoidal waveform. At present, unless first cost is a major consideration, nearly all a.c. bridges employ a valve oscillator as a power source. The advantages of this are that almost any frequency is obtainable, with a sinusoidal waveform, and, where desired, the frequency may be varied. Thus capacity, inductance and h.f. resistance of components may be measured at working frequencies. Where the testing frequency is fixed and its value is not of primary importance, it is customary to employ 1000 c/s. This is advantageous when aural means are employed for balance detection because the sensitivity of the ear is a maximum in this region. Where it is essential that some property of a component shall be measured at supersonic or radio frequencies the valve oscillator is essential as a power source. Thus bridges have been constructed to operate at a frequency as high as 10^7 c/s.

Detectors.

With early a.c. bridges the null-point, or balance, was found by means of earphones. At 50 c/s this is not very satisfactory owing to the relatively low sensitivity of the ear in this region. The average ear has maximum sensitivity at about 3000 c/s and, at the threshold of audibility, the intensity level of a 1000 c/s note must be about 8 decibels above this. However, the ear is still relatively sensitive at 1000 c/s, *i.e.*, at the frequency at which most bridges operate, and earphones still form one of the most sensitive means of detecting the point of balance. However, phones may place an undesirable load on the bridge, or the bridge may operate at supersonic or radio frequencies. In the first of these circumstances the phones may be operated through a valve amplifier when, of course, greater sensitivity will be obtained and the bridge loading will be negligible. If aural means of balance are desired at supersonic frequencies, it is possible to heterodyne the bridge out-of-balance voltage. This, of course, necessitates an additional oscillator, a mixer valve and an intermediate frequency amplifier.

Visual Detection.

When it is desired to detect the balance point by visual means several alternatives are available. If sensitivity and loading of the bridge network are not very important a sensitive rectifier voltmeter may be connected directly to the bridge. For l.f. work the rectifier may be of the copper-oxide type while for h.f. work a full-wave germanium rectifier may be employed. When sensitivity and loading are important a valve voltmeter preceded

by an amplifier may be used.* The necessary frequency characteristics of the amplifier will, of course, depend on the frequency band over which the bridge operates.

As an alternative to a meter, a c.r. tuning indicator may be employed. The output of the bridge amplifier should be rectified and then applied between grid and cathode of the indicator. Since the output voltage of the bridge is zero at the null point, the shadow angle will be a maximum at this point.

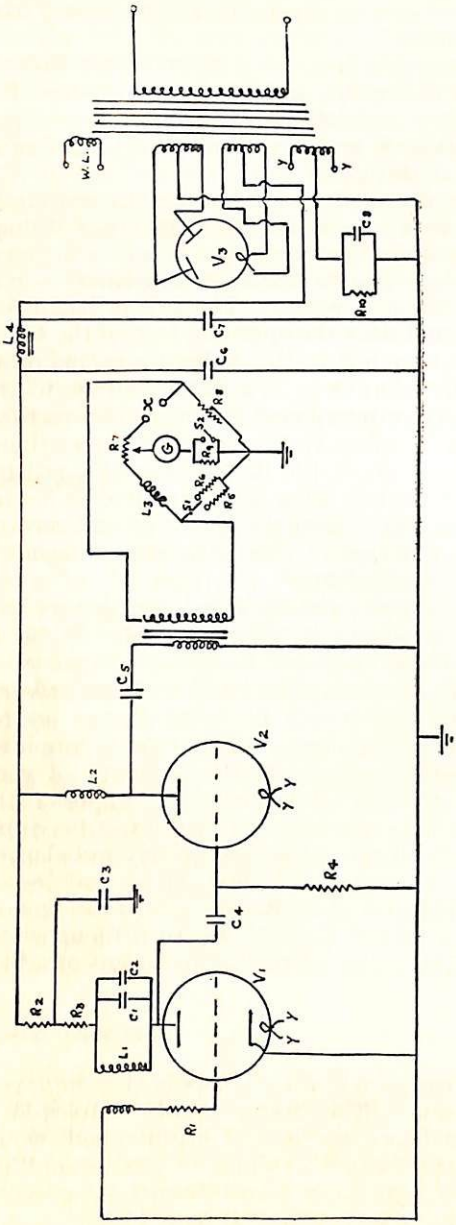
Fig. 54 shows a bridge, designed by the author, to operate at 100 Kc/s. This bridge is used to measure the inductance of dust-cored chokes, employed in suppressor circuits. It has two ranges 0—200 and 0—1000 μH. A master oscillator-amplifier is employed with an output of about 5 watts. The bridge has a low impedance, about 78 ohms, and since the optimum load of the PX25 is 5,500 ohms the bridge is matched to this valve by means of an 8·8 : 1, Radiometal core, transformer. The input voltage to the bridge is 17 volts and balance is determined by means of a rectifier voltmeter. The rectifier is a full-wave Westector, these rectifiers functioning satisfactorily up to about 100 Kc/s. The 47K resistor is normally in series with the meter, being short-circuited to obtain final balance. The shorting switch is of the bias type and returns to the normal position after release. This, of course, safeguards the meter when the bridge is unbalanced.

A further bridge, of the author's design, indicating the application of electronics thereto, is shown by Fig. 55. This bridge is employed for the measurement of power factor and capacity of capacitors at 1000 c/s. The consumption of the bridge network is about 22VA at 110 volts and it will be noted that a master oscillator-push-pull amplifier is employed. This type of amplifier is used because of the heavy loading and the necessity of good waveform. The detector consists of a one-stage amplifier the output of which is fed to a 500 microammeter of the rectifier type. The amplifier gain is 68. With the bridge completely unbalanced the output voltage is 90. To protect the detector R_5 and R_7 are employed, the latter limiting grid current, the former being normally in circuit. Towards the null point R_7 is without effect while, for final adjustment, R_5 is short-circuited by means of a bias switch.

Power Supplies.

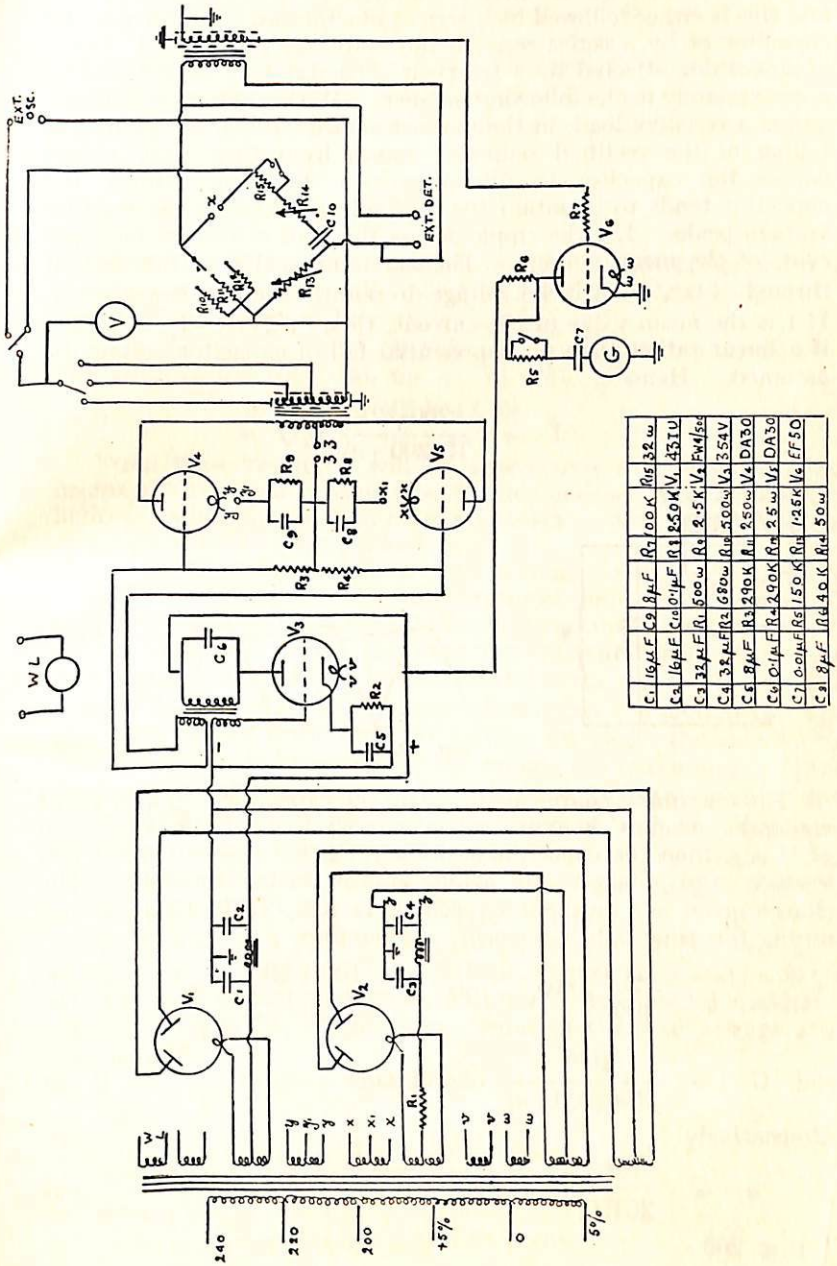
In most electronic instruments a d.c. supply is employed, derived from the a.c. mains. The mains supply employed is invariably single-phase and hence the d.c., if unsmoothed, would possess an intolerably large ripple voltage. The smoothing equipment commonly used takes the forms shown in previous

* *An Improved Amplifier and Null Indicator*, W. R. Thurston, General Radio Experimenter, February, 1948.



R1	3·3K 1W	R10	6·0ω 2W	V1	354V
R2	33K 1W	C1	0·01μF	V2	PX25
R3	15K ½W	C2	0·0042μF	V3	43-IU
R4	1M 1W	C3	0·1μF	L1	100μH
R5	20ω 10W	C4	0·01μF	L2	18·6mH
R6	100ω 5W	C5	0·1μF	L3	202·5μH
R7	100ω 4W	C6	16μF	L4	1·2H
R8	100ω 4W	C7	16μF	G	100μA
R9	47K 1W	C8	0·1μF		

Fig. 54.



C1	16μF	C9	8μF	R1	100K	RG	32 ω
C2	16μF	C10	0·1μF	R11	250K	V1	4S1U
C3	32μF	R1	500 ω	R12	2·5K	V2	FW4/500
C4	32μF	R2	680 ω	R13	500ω	V3	354V
C5	8μF	R3	290K	R14	250ω	V4	DA30
C6	0·1μF	R4	290K	R15	2·5 ω	V5	DA30
C7	0·01μF	R5	150K	R16	125K	V6	EF50
C8	8μF	R6	40K	R17	50 ω		

Fig. 55.

instruments, *i.e.*, the rectifier is followed by a reservoir capacitor and this is either followed by a series inductor and a further parallel capacitor or by a series resistor and stabilizer tube. The degree of smoothing effected by a reservoir capacitor only may be found approximately in the following manner. With a capacitor shunted across a resistive load, in the manner shown by Fig. 56, periodical falling of the rectified semi-sine waves from their peak values causes the capacitor to discharge into the load. Hence the capacitor tends to maintain the d.c. voltage between the rectifier voltage peaks. Let the ripple across the load resistance be a per cent. of the mean voltage. The maximum value of the current through R is $\sqrt{2}V/R$ if the voltage drop in the rectifier is neglected. If I is the mean value of the current, then $(\sqrt{2}V/R - I) = Ia/200$ if a linear rather than an exponential fall of capacitor voltage be assumed. Hence

$$I = \frac{200\sqrt{2}V}{R\,(200+a)}$$

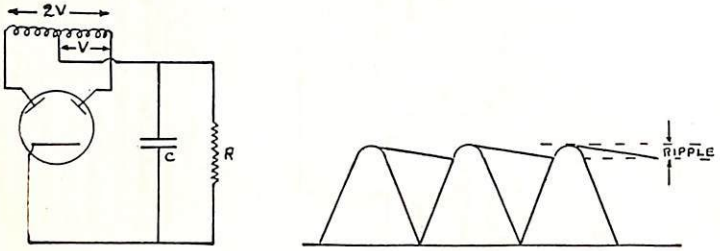


Fig. 56.

The maximum charge acquired by the capacitor is $\sqrt{2}VC \times 10^{-6}$ coulombs, where C is expressed in microfarads. If the frequency of V is f, then the capacitor is discharging for approximately $1/2f$ seconds and charges up again almost instantaneously. The charge given out during $1/2f$ seconds is $100\sqrt{2}V/Rf\,(200+a)$ and during this time falls a per cent. Therefore

$$\frac{a}{100}\sqrt{2}VC \times 10^{-6} = \frac{100\sqrt{2}V}{Rf\,(200+a)}$$

$$\text{and} \quad C = \frac{10^{10}}{Rfa\,(200+a)} \text{ microfarads} \tag{29}$$

Alternatively

$$a = \frac{10^8}{2CRf} \tag{30}$$

if $a \ll 200$

Referring to Fig. 56, if the rising portions of the ripple voltage are regarded as straight lines (as they practically are if the tube drop is considered) then the ripple waveform is approximately triangular. Since the r.m.s. value of a triangular wave is $1/\sqrt{3}$ times the maximum value it follows from (30) that the r.m.s. value of the ripple voltage expressed as a percentage of the mean voltage is

$$\frac{a}{2\sqrt{3}} = \frac{10^8}{4\sqrt{3}\,CRf}$$

the quantity $10^6/4\sqrt{3}\,CRf$ being termed the ripple factor when C is expressed in microfarads. If E is the mean voltage across R, then the regulation of the rectifier is given by

$$E = \sqrt{2}V - aV/200$$

and substituting from (30) for a

$$E = \sqrt{2}V - 10^6\,V/4\,CRf$$
$$\fallingdotseq \sqrt{2}V - 10^6\,I/4\,Cf$$

From these results it will be appreciated that smoothing by means of a reservoir capacitor is effective only when R is relatively high, *i.e.*, when the rectified current is small. Also the advantages of a high value of f are evident.

From (30) it might appear that a could be made indefinitely small by increasing C. In practice upper limits are given for C by valve manufacturers because beyond certain values, depending on the rectifying valve, the capacitor current during the short charging period is likely to be detrimental to the valve. If reference is made to Fig. 56, it will be appreciated that conduction through the valve occurs when the load voltage intersects the sine waves and ceases shortly after the supply voltage is a maximum. If the drop in the valve is ignored, the voltage across the capacitor and the resistor during conduction is $\sqrt{2}V\sin pt$ and the capacitor charging current, being equal to Cdv/dt, is $\sqrt{2}Vp\mathrm{C}\cos pt$. The resistor current is $\sqrt{2}V\sin pt/R$ and hence the valve current is

$$' = \sqrt{2}V\sqrt{p^2C^2 + 1/R^2}\,\sin(pt + \phi)$$

where ϕ = arc tan pCR. This current ceases for $(pt_1 + \phi) = \pi$, or $pt_1 = (\pi - \phi)$. Thus if C = 16μF, R = 2000 ohms, and f = 50 c/s, $\phi \fallingdotseq 84°$ and conduction ceases 6° after V reaches its maximum value. If v is the instantaneous value of the load voltage after conduction

$$C\,dV/dt + v/R = 0$$

or $$v = Ae^{-t/CR}$$

When $t = t_1$, $v = \sqrt{2}V\sin pt_1$, and hence

$$v = \sqrt{2}V\,\sin pt_1 e^{-(t-t_1)/CR}$$

or, since $t_1 = (\pi - \phi)/p$,

$$v = \sqrt{2}V\sin pt_1 e^{-(pt - \pi + \phi)/pCR}$$

When conduction recommences, say, at time t_2

$$\sqrt{2}V\sin pt_2 = \sqrt{2}V\sin(\pi - \phi)\,\epsilon^{-(pt_2 - \pi + \phi)/pCR}$$

or $$\sin pt_2 = \sin(\pi - \phi)\,\epsilon^{-(pt_2 - \pi + \phi)/pCR}$$

which can only be solved graphically. However, solution by this means permits the determination of the peak current through the valve because this is given by $\sqrt{2}V\sqrt{p^2C^2 + 1/R^2}\sin(pt_2 + \phi)$ or, approximately, by $\sqrt{2}VpC\cos pt_2$. Since $\sqrt{2}V \fallingdotseq RI$ the ratio of the peak to mean current is $pCR\cos pt_2$, and, for the quantities previously assumed, is 5·6.

For reasons that will now be evident the reservoir capacitor is usually limited to a recommended value and a further smoothing section or sections added. Considering a section consisting of a series inductor and parallel capacitor, the ripple on the reservoir is approximately the same as when these are absent. Since the reactance of the smoothing capacitor is invariably small compared with that of the inductor and the load, the percentage ripple on this capacitor is aX_c/X_l where X_c and X_l are, respectively, the reactances of the smoothing capacitor and inductor. The ripple on the inductor is approximately the same as that on the reservoir. It will be noted that X_c/X_l is inversely proportional to the square of the ripple frequency and, from (30), it follows that the final ripple on a π section filter, such as that described, varies inversely as the cube of the ripple frequency.

Choke Design.

Since the series inductor, or choke, carries a d.c. current this fact must be considered in design because these chokes, at low frequencies, are built with a laminated iron core. For a choke of this type the author has shown elsewhere* that

$$L = \frac{1{\cdot}25n^2 \times 10^{-8}}{l/a\mu_A + l_g/a_g} \text{ henrys.}$$

where L is the self-induction, n the number of turns, l and a the length and cross-section of the iron circuit, l_g and a_g those of a series air gap and μ_A the incremental permeability of the iron. Since a and a_g are approximately the same, then, writing $l_g = \beta l$,

$$L = \frac{1{\cdot}25n^2 a\mu_A \times 10^{-8}}{l\,(1 + \beta\mu_A)} \qquad (31)$$

If I is the value of the d.c. current in the choke windings

$$1{\cdot}25nI = Hl + \mu H\beta l$$

* *Aircraft Electrical Engineering*, p. 241.

so that $$H = \frac{1{\cdot}25nI}{l\,(1+\beta\mu)}$$

where μ is the ordinary permeability and H is the d.c. magnetizing force. Multiplying (31) by $I/(1+\beta\mu)$ gives

$$\frac{IL}{1+\beta\mu} = \frac{Hna\mu_A \times 10^{-8}}{1+\beta\mu_A} \tag{32}$$

If V_r is the r.m.s. value of the ripple voltage on L, then

$$\triangle H = \frac{1{\cdot}25n\sqrt{2}V_r}{2\pi fLl\,(1+\beta\mu)}$$

$$\text{or} \quad \frac{\triangle H}{H} = \frac{\sqrt{2}V_r}{2\pi fIL} \tag{33}$$

and, from (32)

$$na = \frac{IL\,(1+\beta\mu_A)\times 10^8}{H\mu_A\,(1+\beta\mu)} \tag{34}$$

$$l = \frac{1{\cdot}25nI}{H\,(1+\beta\mu)} \tag{35}$$

The volume of the core is $v = al$ and, from the last two expressions

$$v = \frac{1{\cdot}25I^2L\,(1+\beta\mu_A)\times 10^8}{\mu_A\,(H+\beta B)^2} \tag{36}$$

If all quantities but I^2L are constant in (36), this expression indicates the necessary change in volume when the d.c. current and a.c. inductance are modified.

In order to make practical use of the foregoing results it is necessary to have access to data giving μ_A, μ, B and H for the iron to be employed. Some typical data for 4% silicon iron are given by Fig. 57, and with the aid of these the design of two chokes will be undertaken.

The first choke to be considered is to have an inductance of 50 henrys when carrying 20*m*A, with a ripple voltage of 11 volts of 100 c/s. In order that L shall be practically independent of the value of I, the author finds that β should be equal to about 0·004 and the iron should be worked near its point of maximum permeability. From (33) $\triangle H/H = 0{\cdot}025$. With $H = 1$, $\triangle H = 0{\cdot}025$ and, from Fig. 57, $\mu_A = 430$. Substituting in (34) $na = 24{,}000$, and with $a = 2$ sq. cms., $n = 12{,}000$ turns. From (35) $l = 11{\cdot}5$ cms. On winding the core of Fig. 58 with 12,000 turns of 39G wire, the following results were obtained :—

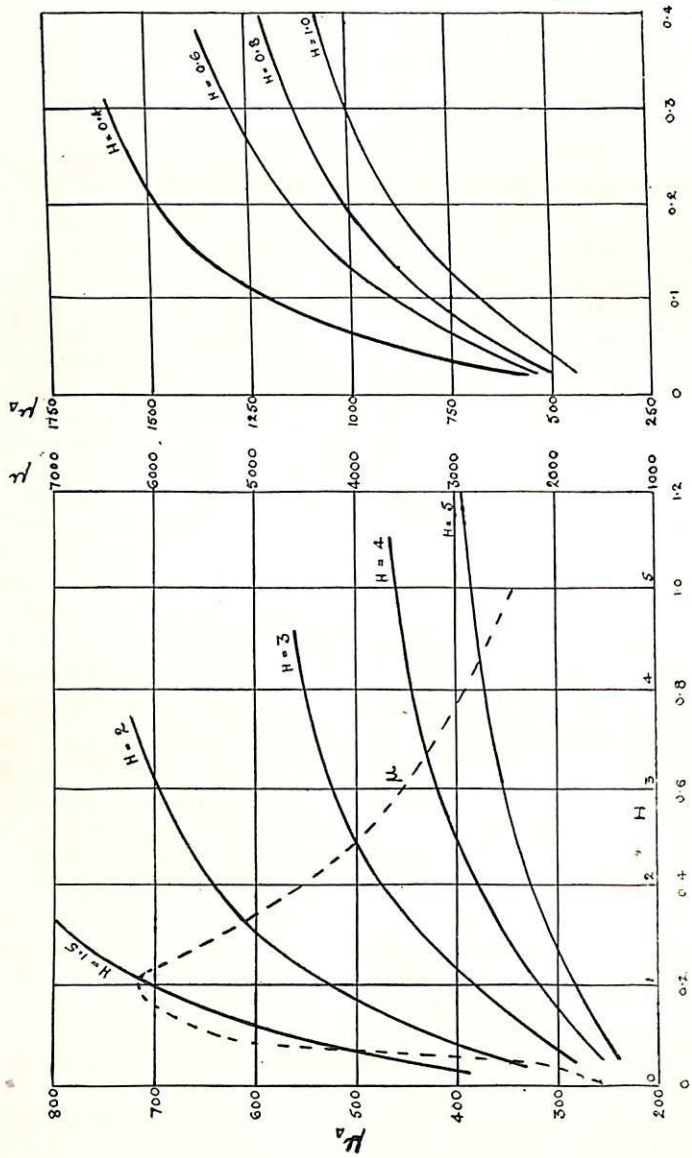


Fig. 57.

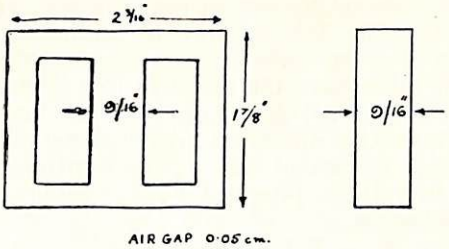


Fig. 58.

Table 2.

mA	*Henrys*
—	59
5	57
10	52·5
15	52·5
20	52·5
28	52·5

The same core was employed for a choke having an inductance of 1 henry when carrying 140 *m*A, with a ripple voltage of 7·5 volts of 100 c/s. With H = 1, $\triangle$H = 0·122 and μ_A = 745. Hence n = 1550 and l = 10 cms. On winding the core with 1550 turns of 30G wire the following results were obtained :—

Table 3.

mA	*Henrys*
0	1·2
50	1·2
100	1·14
140	1·06
200	0·95

Stabilization.

It will have been noted that in the majority of the foregoing instruments stabilization of the H.T. supply is necessary and that in some instances it is desirable or necessary to stabilize the bias and heater supplies. Stabilization may be effected on either

the A.C. or D.C. side of the system but it is the latter side that is usually treated. The H.T. supply is commonly stabilized by means of a gas discharge tube arranged in the manner shown by Fig. 59. Such tubes have the property that a change of current results in a relatively small change in voltage on the tube terminals, *i.e.*, the tube possesses a small dynamic, or slope, resistance. Thus if I is the change in current and R_2 the dynamic resistance, R_2I is the change in voltage, provided R_2 is constant. The equation of the tube voltage is

$$E_3 = E_2 + R_2I_2 = R_3(I_1 - I_2) \tag{37}$$

where E_2 is constant and I_2 is the tube current.

Also $$E_1 = R_1I_1 + E_2 + R_2I_2 \tag{38}$$

where I_1 is the total current through R_1 and E_3 may be termed the burning voltage.

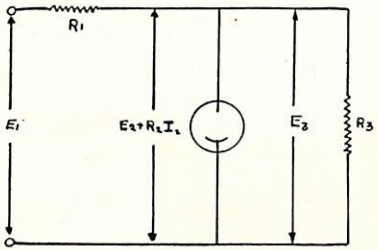


Fig. 59.

From (37) and (38)

$$E_3 = \frac{E_1 + R_1E_2/R_2}{1 + R_1/R_3 + R_1/R_2} \tag{39}$$

and

$$\frac{dE_3}{dE_1} = \frac{1}{1 + R_1/R_3 + R_1/R_2}$$

which, for most practical conditions, may be written as R_2/R_1. Also

$$\frac{dE_3/E_3}{dE_1/E_1} = \frac{R_2E_1}{R_2E_1 + R_1E_2} \tag{40}$$

$$\fallingdotseq R_2E_1/R_1E_2$$

which gives the ratio of the percentage change in E_3 to the percentage change in E_1. It is apparent that the larger R_1 becomes the smaller is the change in the load voltage consequent upon a change in supply voltage. However, for a given current I_1, an increase in R_1 necessitates an increase in E_1 and this tends to set an upper limit to the value of R_1.

In many instruments the load current I_3 remains approximately constant. However, in cases where this varies the change in E_3 consequent upon a change in I_3 is important. Writing $R_3 = E_3/I_3$, substituting in (39) and simplifying

$$E_3 = \frac{R_2E_1 + R_1E_2 - R_1R_2I_3}{R_1 + R_2}$$

and

$$\frac{dE_3}{dI_3} = \frac{-R_1R_2}{R_1 + R_2} \fallingdotseq -R_2 \qquad (41)$$

Efficiency.

The efficiency is given by

$$\eta = \frac{E_3I_3}{E_1I_1} \qquad (42)$$

$$\fallingdotseq \frac{E_2^2R_1}{R_3E_1(E_1 - E_2)}$$

and, owing to the losses in R_1 and the tube, is relatively low.

Design.

A number of different types of stabilizer tubes are available, each type being distinguished by its values of E_2 and R_2, maximum, quiescent and minimum currents, and striking potential. Some values of these quantities for various tubes are given by Table 4.

Table 4.

Type	E_s	E_2	R_2 (ohms)	I_{max} (mA)	I_Q (mA)	I_{min} (mA)
86A1	125	83-87	290	8	4·5	1
4687	130	90-110	250	40	20	10
13201A	135	90-110	80	200	100	15
S130	160	115-135	70	75	—	10
VR75	100	75	—	40	—	5
KD60	85	60-63	—	2·5	—	0·125
G50/1G	90	50	—	0·5	—	2

I_Q is the value of I_2 under normal operating conditions while I_{min} is the minimum value of I_2 for stable operation. Where the voltage of a single tube is inadequate, higher voltages may be obtained by connecting several tubes in series with a common

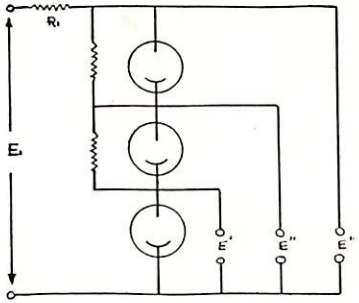


Fig. 60.

series resistor in the manner shown by Fig. 60. By this means a number of different stabilized voltages may be obtained if desired. In order to facilitate striking without the necessity for a high supply voltage a high resistance may be shunted across one or more of the tubes. With a single tube connected to a permanent load we must have

$$E_1 > E_s\,(R_1 + R_2)/R_3 \qquad (43)$$

The design of a stabilizer to supply 40 *m*A at 120 volts will now be considered, the quantity given by (40) to be 0·1. Employing an S130 tube, I_2 will be taken as 35 *m*A.

Hence

$$\frac{E_1 - 120}{R_1} = 0{\cdot}075$$

$$\text{and} \quad \frac{70E_1}{70E_1 + 120R_1} = 0{\cdot}1 \qquad (44)$$

Solving these equations, $E_1 \fallingdotseq 200$ volts, $R_1 \fallingdotseq 1050$ ohms. Substituting in (41) it will be found that $dE_3 = -2{\cdot}65$ volts and this gives the regulation between no-load and full-load. From (43) $E_1 > 216$ volts. However, before the tube strikes the loading is less than 0·075A and with a supply of poor inherent regulation the voltage would be adequate to cause striking. From (42) the efficiency is approximately 31·5%. Practical tests with an S130 and the above values of E_1 and R_1 gave $dE_3 = -3{\cdot}0$ volts and (40) as 0·1.

A.C. Stabilization.

When it is desired to stabilize the heater supply in addition to that of the H.T., several alternatives are available. One is to employ a voltage regulating transformer.* However, in their

* *Voltage Regulating Transformers*, R. O. Lambert, *Electronic Engineering*, February, 1943, p. 384.

simpler forms, suitable for electronic instruments, these transformers suffer from a number of disadvantages such as magnetic saturation, low power factor, waveform distortion, poor regulation, etc. A method of stabilizing the heaters is by means of a barretter. This is placed in series with the heater (or heaters) and will hold the heater current constant over a fairly wide range of input voltages. Thus a BU 30/2 barretter will maintain a current of 0·3A constant for a voltage variation of 1-3 volts across the barretter while a BU 30/50 will maintain the same current constant for a voltage variation of 25-75 volts. Should the heater current be less than that for which the barretter is rated a resistance should be shunted across the heater so that the combined heater and resistance currents equal that of the barretter. It is apparent that the greater the voltage range of the barretter the greater is the stabilization. However, the efficiency of the heater circuit is lowered as the barretter voltage is raised and this may be a limiting factor unless a number of heaters may be operated in series.

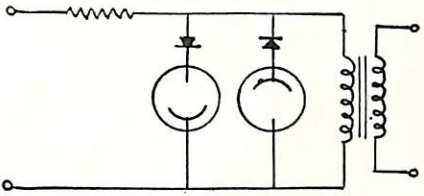


Fig. 61.

It is apparent that when stabilization is effected on the primary side of the mains transformer all the secondary a.c. voltages are stabilized with respect to variations in the supply voltage. As an alternative to a voltage regulating transformer the arrangement of Fig. 61 may be employed. Here it will be noted that the transformer primary is connected across a pair of stabilizer tubes, the latter being connected in opposite manners. When the electrodes of the tube are of similar areas and materials it is possible to employ a single tube only. The author has obtained satisfactory results with a 4687 tube. With different areas and materials two tubes must be employed and so arranged that the glows occur on the surfaces of the cathodes. To obtain a pair of tubes that will operate in the correct manner may be found difficult. Each tube should operate for half a cycle only with the correct electrodes acting as anode and cathode, respectively. To ensure this a half-wave rectifier may be placed in series with each tube, as shown. Alternatively a pair of cold cathode, gas-filled triodes may be employed, such as the 1267 type.

The voltage across the tubes and the transformer primary consists of a pre-striking sinusoidal portion followed by an approxi-

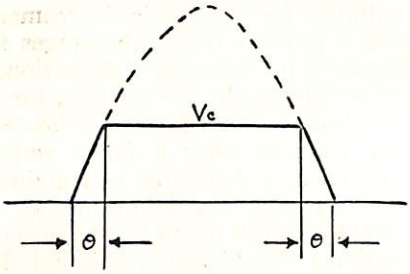


Fig. 62.

mately constant part, V_c, and, finally, a portion similar to that of the initial part of the wave. This is indicated by Fig. 62 and the r.m.s. of the primary voltage is, approximately,

$$V_p = \sqrt{\frac{2.2V^2p}{\pi}\int_0^{\theta/p}\sin^2 ptdt + \frac{p}{\pi}\int_{\theta/p}^{(\pi-\theta)/p} V_c^2\,dt}$$

where $\theta = \arcsin V_c/\sqrt{2}V$.

The value of this expression is

$$\sqrt{\frac{V^2}{\pi}(2\theta - \sin 2\theta) + \frac{V_c^2}{\pi}(\pi - 2\theta)}$$

which is approximately equal to V_c, if θ is small compared with π. Thus if $V = 240$ and $V_c = 100$, then $V_p = 90$ volts. Under these conditions the primary waveform is approximately rectangular and, provided the transformer magnetizing current is low, the secondary load may be referred to the primary side in the usual manner. Then the arguments previously applied to d.c. stabilization are applicable. Of course, in practice, the secondary of the transformer is usually connected to some form of rectifier. Because the voltage applied to the rectifier is rectangular in form, a much smaller reservoir capacitor may be used than when a.c. stabilization is not employed.

GENERAL REFERENCE.

"Electronics," F. G. Spreadbury. (Pitman).

BIBLIOGRAPHY.

"A New Type of Electron-Optical Voltmeter," L. Jacob, J.I.E.E., December, 1944, p. 512.

"A Valve Ammeter," H. E. M. Barlow, J.I.E.E., Vol. 77, p. 612.

"An Electronic Pressure Gauge," H. A. Prime and T. J. Chi, *The Engineer*, March 25th, 1949, p. 320.

"The Evolution of the Quartz Crystal Clock," W. H. Marrison, *Bell System Technical Journal*, July, 1948, p. 510.

"An Electronic Insulation Tester," J. S. Forrest, J.I.E.E., Vol. 89, Pt. II, p. 288.

"An Electrical Moisture Meter," C. F. Brockelsby, J. Sci. Insts., Dec., 1945, p. 243.

"Improved Cathode-Ray Tubes with Metal-Backed Luminescent Screens," D. W. Epstein and L. Pensak, *R.C.A. Review*, March, 1946, p. 5.

"A New Fuel Gauge for Aircraft," *The Engineer*, March 1st, 1946, p. 204.

"The Stabilistor," A. B. H. Walker, *Wireless World*, November, 1944, p. 339.

"Electronics in Measurements," R. J. Kryter, J.A.I.E.E., Jan. 1947, p. 31.

"Impedance Bridges for the Megacycle Range," H. T. Wilhelm, *Bell System Technical Journal*, Sept. 1952, p. 999.

"Technique of Trustworthy Valves," E. G. Rowe, Proc. I.R.E., Oct., 1952, p. 1166.

"Concerning the Reliability of Electron Tubes," M. A. Acheson and E. M. McElwee, Proc. I.R.E., Oct. 1952, p. 1204.

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